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AN OPERATIONAL AMPLIFIER PRIMER UNEXPURGATED

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UNDERWATER SYSTEMS DEPARTMENT

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This report discusses the use of integrated circuit operational amplifiers, the most common linear (analog) integrated circuit. Both the idealistic theory and the practical limitations of actual devices are discussed. Most common basic circuits are given. Some examples of circuits in actual use are included for illustration. This report is an update of a previous report issued in 1980.

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FOREWORD

This report is an update of a previous report written in 1980.[1] That report followed a 1972 report which followed an internal publication which was published in 1970. This report contains all the material from the previous reports plus additional subject matter. All previous notes are superseded by this report and may be destroyed.

The purpose of this report is to point out the uses and limitations of operational amplifiers to those unfamiliar with them and perhaps extend the knowledge of those to whom they are already familiar. The subject is presented from an introductory, practical point of view and is not an exhaustive treatise on the subject by any means; neither is it a "cookbook" of ready-to-go-circuits. These can often be found in manufacturers' application notes; but, if one blindly applies someone else's circuit to his own job without understanding the principles and limitations, often the circuit does not do the job. Examples of actual devices are limited to monolithic and semi-monolithic integrated circuit types, as these are smallest, cheapest and, hence, most common. Specific types and manufacturers are given only as examples and are not necessarily good or bad.

It is interesting to note what has changed since 1970, as progress in integrated circuits has been fast and furious. The basic theory of the original report still holds true. The basic circuits are still useful, although some entirely new circuits have since been published. On the other hand, the majority of the integrated circuit types originally listed are now obsolete! Some of the predictions made along the way were accurate, some missed the mark; and some of the developments were not foreseen at all.

The original report was written because there was a remarkable shortage of books on the subject. Since then the reverse situation has occurred; a rash of books appeared of varying merit. The author's preferences are as follows. Reference 2 is very thorough, consequently quite long and involved, and it is probably the best general reference. In addition, there are several "follow-on" books in the series which delve even further into the intricacies of the subject. Reference 3 is the opposite extreme, being intended for persons not familiar with electronics at all. Integrated circuit manufacturers often publish helpful application notes, for example, References 4 through 7. Indeed, the specification sheets for individual types frequently include circuit applications.

Why then is this report still necessary? Well, it still seems to fill a gap. It is a good concise summary. Many books go into great detail on such subjects as the design of monolithic integrated circuits, which most users do

not care about. Another pitfall is a painfully thorough description of the author's favorite circuit that no one will ever again use. Many books are out-of-date when written. Some authors have clearly never been out of the academy; they analyze irrelevancies to death but miss the important points. Also, two companion reports to this one have now been published.[8,9]

The three reports serve as textbooks for a course that has been taught for several years. Much of the extra material presented in the teaching of the course has been included in this edition to make it more complete. This has lengthened the report significantly, but the previous editions were probably oversimplified. The serious student or circuit builder in many cases did not have adequate information. The reader should not be awed by the length of the report. Entire sections may be skipped if not applicable to the reader's requirements.

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INTRODUCTION

The operational amplifier, abbreviated op-amp or simply OA, is receiving continuously more widespread use for several reasons. First, it is probably the most versatile analog integrated circuit available, accounting for about half of all analog sales. Some of its uses are: summer, integrator, amplifier, buffer, converter, complementer, clipper, filter, thresholder, oscillator, regulator, and even logic. Second, through silicon integrated circuit (IC) technology, it is now available in a .05 in^3 package for less than $1.00. Thus, it can be used almost as indiscriminately as a transistor. The past few years have witnessed significant improvements in most op-amp parameters. Actual devices are closer to ideal than ever before. Consequently, they are finding an even wider range of application.

HISTORY

Digital computers have become so ubiquitous that few recall that the first electronic computers were analog. That is, they analyzed a system, say mechanical, by setting up an electronic system governed by the same equations. Behavior of the electronic "analogue" could be observed and adjusted faster, easier, and cheaper than the mechanical, as a potentiometer ("pot") could be adjusted rather than a new part machined. And the results of total failure were usually far less spectacular and costly with electronics.

Analog computers worked. For general-purpose and high-precision work, they were rapidly replaced by digital, once the initial problems of the latter were worked out. However, some of the techniques and components were useful enough that they were retained along with the name analog (which had been shortened slightly). We owe a great deal to the early pioneers such as Black of Bell Labs, who showed the importance of making circuit behavior depend on passive feedback components only and not on the active devices, and Philbrick of the company that still bears his name, who figured out ways of making decent op-amps with vacuum tubes. (Transistors hadn't been invented yet.)

The term "linear" is often substituted, but nearly half of analog is not linear at all, so this report will use "analog." Analog will be around for a long time yet, and its chief component will be the op-amp.

DEFINITION

The simplest possible representation of an op-amp is shown in Figure 1a. The output voltage (signal), referenced to ground, is equal to the input voltage multiplied by some gain (A) which is much, much greater than one and inverted, indicated by the minus sign. This representation tells only half the story, however. Op-amps almost invariably have two inputs, as shown in Figure 1b, termed negative and positive, or inverting and noninverting. The output voltage is equal to the gain times the voltage difference between the two inputs. A third type, shown in Figure 1c, sometimes appears which has two outputs, where one output is the negative of the other. This type is seldom used, as the type of Figure 1b usually will suffice by virtue of having noninverting and inverting inputs. In any event, a type c can be realized with two of type b. Only type b will be discussed in this paper. If only one input is shown, assume that it is the inverting input and that the noninverting input is grounded.

For an ideal op-amp, the gain (A) is infinite and is independent of frequency, the input impedance is infinite so it places no load on the source, and the output impedance is zero so the output voltage is unaffected by any load placed on it. In the formulas to follow, an ideal op-amp is assumed. Actual op-amps, of course, fall short of all these ideal characteristics but, in most well-designed circuits, the limitations can be ignored and the ideal equations hold to a reasonable approximation. In fact, in some circuits, some of the assumptions will turn out to not even be necessary. The actual limitations will be discussed toward the end of the report.

The op-amp is defined as a device having high voltage gain, but it will not be useful unless it also has high current gain. Having both implies high power gain, which is really that most fundamental aspect, although it is seldom mentioned. Basic physics says this power gained must come from somewhere, so we will always have at least one power supply, often not shown in the circuit. Typically there is an equal pair, one positive and one negative. That is assumed in this report unless mentioned otherwise.

NONLINEAR CIRCUITS

The usefulness of the op-amp can most easily be shown by examples. The simplest possible circuit is a clipper, shown in Figure 2a. If the input voltage waveform is positive, the high gain of the op-amp will drive its output into positive saturation, and the output voltage will be limited by the output swing. Similarly for a negative input voltage, the output will be negative. Thus, the output voltage indicates simply the polarity of the input voltage. If inversion is desired, the two op-amp inputs may be interchanged (i.e., ground the noninverting input and place the signal on the inverting input); the output will be positive for negative input and vice-versa. Input impedance is high and output impedance low due to the characteristics of the op-amp.

Note that the op-amp is really performing the function of a comparator here. In theory an ideal op-amp also makes an ideal comparator. In practice they are quite different, and the reader may feel more comfortable substituting "comparator" for "op-amp" in this section. The two are close enough that the same symbol is used, so Figures 2 and 3 are good either way. This report uses the common convention that "V" denotes a fixed (dc) voltage, while "e" denotes a signal voltage that is normally changing (ac) but could be fixed.

An extension of this circuit, the threshold detector, is shown in Figure 2b. The inverting input is connected to a reference voltage instead of ground. For this circuit the output indicates whether the input is more positive or more negative than the reference voltage. The threshold voltage may be negative; the equation still holds. The output cares nothing about either the input or the reference being positive or negative, only which one is <u>more</u> positive. This is an important consequence of the differential input. Note that ground does not appear in the circuit. It is simply a definition, i.e., zero volts. The output function may be inverted, if desired, by interchanging the inputs as with the clipper.

If the second input is another signal instead of a dc voltage, we have a comparator, shown in Figure 2c. The output indicates simply which of the two inputs is more positive. The clipper and threshold detector can be thought of as special cases of the comparator. Indeed, many of the circuits presented here will be special cases of others, but the general cases are frequently complex and obscure the special uses, so the simpler cases will be presented first.

A common use of this circuit is a clipper where the signal ground is different from the power supply ground, as shown in Figure 3a. A low-level signal may be generated in a different piece of equipment some distance away and transmitted through a coaxial cable or twisted pair of wires. If two equipments do not have a good common ground, a voltage difference may exist between their two chassis. This "hum" voltage will appear as a false signal added to the true signal; it will also appear on the cable shield. If the cable shield is used as a "signal ground," the op-amp will recognize only the difference, which is the true signal. The power supplies may be referenced to either ground. The op-amp will reject noise on the power supplies.

Sometimes noise is induced directly on an unshielded lead, such as high frequency pickup from a nearby digital circuit. In an ordinary clipper, this may cause multiple zero crossings at each true zero crossing. This might confuse the following circuitry; for instance, a zero crossing counter. The circuit of Figure 3b, termed a hysteresis clipper or hysteresis trigger or Schmitt trigger, is useful in such a situation. A fraction of the output voltage is fed back to the noninverting input, termed "positive feedback." When the output starts to switch, regenerative action occurs and it "helps itself" until the output is saturated. The circuit effectively changes its own threshold voltage whenever it switches. When the input rises, the output does not switch at the zero crossing as with normal clipper, but at a positive threshold indicated by the positive dashed line in the figure. When the output switches, it lowers the threshold to the negative line, and the output will not switch back until the input falls below this point. The trigger points are given by the formula shown, where $\pm V$ is the output voltage swing.

The total hysteresis is the difference between the two thresholds. A biasing resistor may be added to move both thresholds positive or negative if asymmetry about zero is desired. In some applications, a capacitor may be added to or used in place of the feedback resistor to eliminate dc hysteresis. This ensures that the output will switch cleanly and then be insensitive to multiple zero crossings caused by noise for a short time, but the noninverting input will settle back near its original state before the next true zero crossing occurs. For ac hysteresis, the approximate signal frequency must be known. If an inversion is not desired, the input and ground may again be interchanged. However, the input impedance for this case is no longer high and, in fact, is nonlinear, as the output affects the input. No simple input-output equation may be written for the hysteresis trigger as the circuit possesses memory.

The somewhat unusual use of an op-amp as a logic element is indicated in Figure 3c. The circuit gives a "go" indication only if all the listed conditions are met. Operation is as follows: the noninverting input is biased slightly positive, so the output is positive if all inputs are normal. If any input goes into its forbidden range, one of the diodes on one of the two buses conducts, switching the output to the "no-go" state. The versatility derives from the op-amp having both noninverting and inverting inputs, whereas a normal digital IC is sensitive only to signals of one polarity. Thus, a single op-amp with diodes, resistors, and a capacitor can provide a rather sophisticated sensing function, replacing several comparators and some digital logic. Hysteresis may be added to this circuit also.

LINEAR CIRCUITS

The reader should note that all circuits discussed up to this point are nonlinear; their outputs are binary, i.e, either on or off. Most of the information has been lost. A given output could result from more than one possible input; there is no "going back." Their usefulness is limited; they are usually not included at all in op-amp books. A larger and more common class is that of linear circuits, where the output amplitude is proportional to that of the input(s). Also, note that none of the previous circuits had negative feedback. Due to the very high gain of the op-amp, negative feedback must be used for stabilization in all linear circuits.

The simplest possible linear circuit is the voltage-follower shown in Figure 4a. The output "follows" the input, so that the gain is +1. There are two general methods for analyzing the behavior of the linear op-amp circuits. The first is the "brute force" method of solving the general circuit equations with all the op-amp parameters and then simplifying the result by approximations until it is manageable. This is usually the hard way. An easier way is to, first, assume that the circuit works. Because of the very high gain of the op-amp, if the output is in its linear range, its two inputs must be at very nearly the same voltage. Using this fact and the assumption that the inputs draw no current, determine if the voltages and currents add up properly according to standard circuit equations, assuming the op-amp is giving the desired output. (If the probable or desired output is unknown,

determine what the output voltage must be for a given input to satisfy the circuit equations and keep the voltage across the inputs at zero.)

Second, if the voltages and currents add up properly, then determine if there is a negative feedback path to stabilize signal (ac) gain and a negative feedback path to stabilize (dc) bias (may be the same path) and that there is no positive feedback path giving a loop gain greater than 1 to cause latch-up or oscillation. In frequency selective circuits, the latter condition becomes: at any frequency where the loop phase shift adds up to a multiple of 360 degrees, the loop gain must be less than 1 for stability.

Lastly, both op-amp inputs must have a dc path to ground. This is guaranteed at the inverting terminal by the second condition above; the noninverting terminal cannot be completely isolated by capacitors. If there is an open input, as in Figure 4a, the driving source must be able to supply the dc bias.

Using this method of analysis, the operation for the follower circuit is apparent. The output voltage is the same as the voltage at the inverting input, so it must remain equal to the voltage at the noninverting input to keep the input difference voltage zero. The feedback is indeed negative, satisfying the second requirement. It should be noted that no parameters of the op-amp itself appear in the overall equation, i.e., the circuit performance does not depend on the op-amp used. This will be true in all the examples. This is, of course, only an approximation, but for most circuits if the proper type of op-amp is selected, it is a very good approximation. If the effect of the op-amp parameters must be determined, the simpler analysis method usually may still be used by determining the ideal response and then working backward and making a small or "first order" error correction to the equation.

Using the correction method, the actual input and output impedances of the follower circuit may be determined. The input impedance of the circuit is high, typically in the megohm range, as the input impedance of the op-amp is high. It is effectively increased by the feedback as, even for large input signals, little voltage is developed across the two op-amp inputs, so there is little change in current. Analysis would show that differential input impedance, the major component of the op-amp input impedance, gets multiplied by the gain around the feedback loop, exactly equal to the op-amp gain in this case. This is the first of three typical input conditions that will be encountered. The circuit output impedance is low, as the op-amp impedance is low. It is lowered further by the feedback, as any small change induced in the output voltage by loading causes an "error" voltage at the input which, in turn, causes a large output current to flow opposing the change. The actual value is the op-amp output impedance divided by the loop gain. This typically calculates to a fraction of an ohm, but this does not mean that the op-amp can drive a one-ohm load. An op-amp typically can supply only a few milliamps current, so its output would go into current-limiting. Any nonlinearity can render the transfer equation invalid, as it was derived assuming linear operation. This will be true of all the linear circuits shown and, hence, will not be mentioned again. Similarly, since the voltage between the two op-amp inputs is not precisely zero, the actual output voltage is equal to the input voltage plus the op-amp input offset voltage and noise voltage, typically in the millivolt range. (These terms

will be defined later.) Op-amp input current, offset current, and noise current will have no effect here as there are no resistors in this circuit; however, if the source has a finite impedance, it will be affected. The output voltage will be reduced by the reciprocal of the op-amp gain, but this amounts to typically 0.001% at low frequencies. Similarly, loading the output reduces the op-amp gain, possibly increasing the loss to 0.002%, which is usually insignificant.

The next circuit is the inverter, shown in Figure 4b. If the two resistors are equal, the output voltage must clearly be equal and opposite to the input voltage to maintain the negative op-amp input at ground, again using the "simple" analysis method. If R_1 and R_2, termed the input and feedback resistor, respectively, are unequal, the gain is given by the formula shown, where the minus sign indicates the inversion. The inverting op-amp input is a "virtual ground"; although it is not connected to ground, it is held very close to ground by the circuit feedback. Remembering this simplifies analysis of many op-amp circuits. The current through all components connected to the inverting input must sum to zero, and each current may be calculated as if the component were connected to the noninverting input, ground in this case. Clearly then, the input impedance of this circuit is equal to R_1. This exemplifies a second type. The input impedance is not infinite, but it is a constant determined by the passive circuit components. The ground used at the noninverting input should be the signal ground, if one is available, to reduce noise in the same manner as the clipper of Figure 3a. (Complete noise rejection is discussed later.) Note that here the op-amp input currents do matter, limiting the maximum resistance values that may be used. To supply a current into the op-amp input, the output voltage must be higher than the input by an amount equal to the product of the feedback resistor and the input current. The "offset" can be improved about an order of magnitude by matching the dc resistances at the two op-amp inputs, in this case by adding a resistor between the noninverting input and ground equal to the parallel combination of the input and feedback resistors. The output voltage error then becomes the product of this resistance and the offset current, defined as the difference between the two input currents. If the two input currents were perfectly matched, they would contribute no error to the output voltage. There would still be an output error equal to the product of the mismatch (difference) between the resistances at the two inputs and the input current, and another equal to the input voltage mismatch (offset) multiplied by the gain from that point, given by the formula in the next example. Note that for the inverter, all feedback and bias considerations are automatically met, even if the stage is ac coupled by adding a capacitor in series with the input resistor.

If gain is desired but the inversion is not, a combination of these two circuits gives the noninverting amplifier of Figure 4c. Input impedance is once again infinite. Gain is given by the formula, which resembles that for the inverter. In fact, there is no reason that the two inputs cannot be used at the same time.

Figure 5a shows a difference amplifier; a voltage divider has been added to the noninverting input to vary its gain. It should be noted from the formula that there are two inherent restrictions with this circuit. First, the gain from the inverting input may be chosen independently of the gain from the noninverting input, but not vice-versa. Second, the gain from the

noninverting input can exceed the gain from the negative input only by one. If these restrictions are a problem, they can always be alleviated by the addition of another amplifier or voltage divider. Note that the input impedance at the inverting input of the circuit is not passive because the current required is affected by the voltage at the noninverting input and, in fact, may be negative to that required by a passive load. This constitutes the third type of input. This nonlinear, undefined input impedance is somewhat undesirable, but it will usually suffice to say that the input impedance is "on the order of" the input resistance. If the signal is coming from the output of an op-amp with less than an ohm impedance and the input resistor is 100K ohm, it doesn't matter whether the latter is positive, negative, nonlinear, or whatever.

The multiple-input circuits may be easily analyzed by the technique of superposition, which holds for all linear circuits. The output response to any individual input is found by setting all the other inputs to zero (ground) and calculating the response in the usual way. The output response when all inputs are used is then given by the sum of the individual responses that have been calculated. The reader should note that the first three linear circuits (Figure 4) can all be derived from this one (Figure 5a) by choosing the right value for each resistor (including zero and infinity) and thinking of ground as an input of zero volts.

An important special case occurs when the gains for the two inputs are set equal. The amplifier then recognizes only the voltage difference between the two inputs. This is important in applications such as strain-gauge bridges or balanced-pair transmission lines where the signal is the small difference between two inputs, each having a large "common mode" dc bias or noise. The two input impedances are mismatched, but high-impedance inputs may be obtained by adding a voltage-follower to each input; this configuration is termed an "instrumentation amplifier." Figure 5c shows this with the addition of a handy gain-adjust network where (differential) gain is set by a single resistor. Note that the other resistors must be matched, at least by pairs, for the common-mode gain to be zero.

If the sum of two inputs is desired, addition of a second input resistor gives the summing inverter shown in Figure 5c. Since the inverting input of the op-amp is a virtual ground, each input sees as input impedance only its own input resistor. Each input resistor supplies a current which must flow through the feedback resistor (the op-amp input takes virtually no current), so that the gains from the two inputs are readily calculated. As shown in the formula, each input basically ignores the other. Note that the two gains may be chosen independently, but then the input impedances may not. This summing method may also be applied to the nonlinear circuits; if the feedback resistor is omitted from Figure 5c, we have a summing, inverting clipper.

Similarly, more than one noninverting input may be applied. Figure 5d shows a general amplifier with multiple inverting and multiple noninverting inputs. In theory any input may be made to have any gain. The equation appears rather complex, but it becomes simple if one can only remember where each term of the expression came from, i.e., what it represents. Again, all previous linear circuits can be derived from this one. However, the complexity of the equation shows the reason for starting with the simple cases and working up.

FREQUENCY-SELECTIVE CIRCUITS

The behavior of all circuits discussed so far has been independent of frequency, at least within the performance capabilities of the op-amps. A more general class is that of frequency selective circuits. Here again it is assumed that the frequency characteristics are determined solely by the passive components and not by the characteristics of the op-amp. (There are some circuits which specifically use the characteristics of the op-amp; these generally create more problems than they solve.)

The simplest and most familiar of this class is the integrator shown in Figure 6a. The output voltage is proportional to the integral of the input voltage, with an inversion, assuming the output was initially at zero. To analyze the response of the circuit, assume the output is initially at zero and a positive step is applied to the input. The inverting input of the op-amp is a virtual ground, so a constant current flow is determined by the resistor. Current through a capacitor is proportional to the time rate of change of the voltage, so to balance the input current and maintain the op-amp inverting input at ground, the output voltage must steadily become more negative, giving the ramp shown. In terms of the generalized frequency "S" used in transient analysis (Laplace transforms), the circuit response is $-1/RCS$. If only periodic waveforms (sine waves, etc.) are of interest, the response simplifies to $-1/RCj\omega$, where ω is the radian frequency, equal to $2\pi f$ (f in hertz) and $j = \sqrt{-1}$, denoting 90 degrees phase advance. In other words, for a sine wave input, the output will be a sine wave whose amplitude is inversely proportional to frequency and whose phase is advanced 90 degrees with respect to the input (90-degree lag from the integration and 180 degrees from the inversion). In terms of impedance, the capacitor in the negative feedback becomes a lower impedance at higher frequency so the overall gain of the circuit goes down with increasing frequency. The product RC is sometimes abbreviated τ, called the time constant of the integrator, analogous to the time constant of an RC network; $1/\tau$ is sometimes called gain since it is a multiplicative constant. If two inputs are desired, a second input resistor may be added as with the summing inverter.

The reader should note that the response of the integrator at zero frequency (dc) is infinite. This problem has two practical manifestations. The first is that for zero input the output need not be zero, depending on the initial charge on the capacitor, so the integrator usually is preset to zero before it is to be used (initialized). The second is that there is no dc negative feedback path for stabilization, so even if the integrator is preset, it will "drift" off slowly due to the input current of the op-amp. Therefore, the output must be held to zero until it is to be used. However, if the integrator is part of a larger system that does provide a dc negative feedback around the integrator, this precaution may be unnecessary (shown in a later example).

A noninverting input may be added as shown in Figure 6b. Response may be analyzed as follows. If a positive step is applied at IN_1, response will be identical to the original circuit of Figure 6a. If a step is applied to both inputs, the output response will be zero, as both the noninverting and

inverting inputs of the op-amp will charge up at the same rate and the difference between them will remain zero. Therefore, the response to the second input must be the negative of the response to the first input. Actually, the capacitors for the two paths need not be the same as long as the RC product is kept the same (τ again).

A circuit complementary to the integrator, the differentiator, may be constructed as shown in Figure 6c. The indicated response may be found by similar analysis. The response of this circuit is zero for dc (indicating dc stability) but infinite for infinite frequency, which can cause the following practical problem: sometimes the desired signal has low-amplitude, high frequency noise riding on it. This noise would normally not cause serious problems but, if the signal is put through a differentiator, this noise may be amplified to unreasonable proportions, perhaps even saturating the amplifier output. The high frequency response can be limited by placing a small resistor in series with the input capacitor. This resistor may have an additional advantage, as the input impedance of this circuit is capacitive, which may cause problems with the driving circuit. Also, some op-amps are simply not stable in the pure differentiator configuration, as will be discussed later; the resistor helps. A noninverting input may also be added to the differentiator as shown in Figure 6d.

Similarly, an integrator may be made dc stable by placing a large resistance in parallel with the feedback capacitor, making the dc gain finite. If the feedback resistor is at least 100 times greater than the input resistor, response is still close to that of a pure integrator, typically within 1%. This circuit is called an averager. A much rarer high frequency problem can also occur with the integrator circuit. High frequency noise would not appear to be a problem with a "low-pass" circuit but, if the noise is fast enough and of large amplitude, it may drive the amplifier out of its linear range and spurious outputs may occur. One possible "fix" for the integrator is a capacitor from the inverting input of the op-amp to ground. This has little effect if the op-amp is in its linear range since the negative input is a virtual ground, but it prevents a rapid voltage change at the op-amp input.

FILTERS

While the above circuits are often used for computation or wave shaping, they are of limited use for frequency selectivity, as they possess different gain for every frequency. Usually, a filter is desired which passes part of the frequency spectrum with uniform gain and rejects the rest. Such an ideal filter is naturally impossible to build. However, there is a useful, buildable class of filters where the amplitude response is fairly uniform or "flat" across a band of frequencies, begins to drop off at the edges, and drops off progressively further at frequencies further away from the desired "passband."

Op-amps are very useful in building active filters. Active filters are frequently preferable over passive filters because they eliminate the need for inductors. A very large number of active filters have been designed.

Reference 9 discusses these in considerable detail. However, the previous report[1] contained a number of filter circuits, which will be repeated here for completeness. The formats of References 1 and 9 are slightly different; the format of Reference 1 will be repeated here. The first example is a low-pass filter, shown in Figure 7a. Operation is as follows. At low frequencies the capacitors have little effect and the gain of the circuit is 1, determined by the negative feedback loop, independent of frequency. At high frequencies the feedback has little effect because of attenuation through the two RC sections, and the response falls off at the rate of 12 dB/octave (equivalent to 40 dB/decade) due simply to the two RC sections. Near the breakpoint frequency between the pass region and the reject region, positive feedback boosts the response to make the breakpoint much sharper than for a simple double RC.

The values shown give a Butterworth, or maximally flat, response characteristic. The properties of the Butterworth filter are that the response is very flat at the bottom end and decreases monotonically with increasing frequency. The transfer function is given in the figure. The magnitude of the transfer function, indicating amplitude response, is also shown. The $\sqrt{1+\omega^4 (RC)^4}$ term shows that the response stays flat until ω is quite near $1/RC$, and soon thereafter drops off as $1/\omega^2$. At $\omega = 1/RC$, the response is down 3 dB. This is the half-power point and is usually taken to be the cutoff frequency of a filter. If some frequency peaking can be tolerated, the dropoff near the breakpoint can be made sharper by increasing the positive feedback slightly. One form of this would be a Tschebycheff filter, which could be obtained with different component values. In fact, any other single pole-pair filter (Bessel, etc.) can be obtained with this circuit.[9] Also, 3-pole and higher order filters may be realized with a single op-amp but component values frequently become unreasonable, so it is usually better to build the filter up from single pole-pair sections. It should be noted that the input assumes the source impedance is zero, or at least low compared to the input resistor. Otherwise, finite source resistance is compensated for in this filter by making the input resistor smaller by the amount of the source impedance.

For a high-pass filter, a circuit complementary to that of Figure 7a is shown in Figure 7b. These two filters are useful for systems requiring exactly unity gain in the passband (0 dB insertion loss). Sometimes gain is desired since, when part of the frequency spectrum is removed, the amplitude of the remaining signal is normally smaller than that of the original. The op-amp can also be used to provide the gain, and many articles have been written on the subject, too numerous to quote. The most difficult way to add gain is to recalculate the component values for nonunity gain. This author has pointed out that the circuit may be modified without recalculating by using Thevenin equivalents.[10] Figure 7c shows the Butterworth high-pass for an arbitrary gain. The transfer function is the same except for the gain factor G. It is interesting to note that making all resistors equal gives a Butterworth filter having a gain of two. The conversion for the low-pass requires a capactive divider (positive feedback side only); this places a capacitive load on the op-amp which can cause problems with some op-amps.

Wilbur[11] has pointed out that the filter component values may be left unchanged if the positive feedback is taken not from the op-amp output but from the inverting input, which "follows" the noninverting input, as long as

the impedance of the negative feedback divider is much, much lower than the impedance of the filter components. However, that is usually not the case; in fact, the two are often made equal, as will be seen later. This type of approximation could be used to eliminate an op-amp in many circuits but, with four op-amps in a package for a dollar, this is usually false economy. This report includes all op-amps so impedances are not critical. Reference 9 points out that, on the other hand, the filter component values may be standardized and the gain used to determine the filter characteristic. Figure 7d is an approximate circuit useful for the high-gain case, taking advantage of the fact that the impedance of the positive feedback series element is large compared to that of the shunt element.

If a wide bandpass circuit is desired, the signal can be passed through a low-pass to remove the high frequencies and then a high-pass to remove the low frequencies, or vice versa. If a wide band-reject is desired, the signal can be fed to both a high-pass and a low-pass and the outputs summed with a summing amplifier. The terms "wide" and "narrow" are, of course, not precise, but here mean roughly more or less than an octave (factor of 2 in frequency).

If a narrow bandpass is desired, the circuit of Figure 8a may be used. At low frequencies, a negative feedback path exists through the resistors and the gain is about unity. Similarly, at high frequencies a negative feedback path exists through C_1 and the gain is also unity. Near the center frequency, the effect of both feedback paths is reduced, so gain is high. The circuit "Q" (defined as the 3 dB bandwidth divided into center frequency) is given by the formula shown. The center frequency is also given. Gain at the center frequency is a function of Q as shown. Center frequency may be adjusted somewhat by making one resistor variable.

The bandpass filter has a gain of one at very high and low frequencies rather than zero, which can be a nuisance in some applications. The filter may be modified slightly as in Figure 8b to avoid this. Bringing the signal in at the op-amp inverting input makes the circuit an integrator at high frequencies. A coupling capacitor is added so the circuit acts as a differentiator at low frequencies. The same formulas apply approximately except that the center frequency gain becomes a more complicated expression. The high gain at the center frequency is reduced by making the input resistance large.

Another variation is shown in Figure 8c. The capacitors and resistors in the bridged-tee have been interchanged so the Q is determined by a ratio of resistances. (When redrawn as shown, the circuit is not easily recognizable as a bridged-tee.) Since the shunt resistor is small for high Q, the input current is injected at that point making the input resistance a reasonable value. This version is preferable over Figure 8a because two resistors having grossly different values may still be made from the same material and their values will track, whereas two grossly different capacitors almost certainly will not. It has the disadvantage that the op-amp sees a load to ground of a capacitance in series with a small resistance, and it also has a near 90-degree lag feedback path to the inverting input consisting of a large resistance in series and a capacitor

to ground; either condition can make some op-amps unstable. Here the Q may be varied by making one resistor variable; however, this changes the center frequency so the other resistor must also be variable to compensate.

Frequently, signal processing systems require filters of octave bandwidth, the borderline between wide and narrow bandpasses. One way to do this is to "staggertune" two (or more) narrow-band stages, as shown in Figure 8d. This filter gives steep slopes at the band edges (24 dB in the first octave outside the band) at the expense of high ripple (about 3 dB). This is a rather crude filter, having more ripple than would be tolerated in most applications. Ripple may be reduced by lowering the Q of the stages, which makes the slopes less steep, or else by adding more stages (poles), for example, a third stage centered in the band where the valley in the response is (see Reference 9).

A complement of the narrow bandpass, used to reject a single frequency, is the narrow band-reject, or notch, shown in Figure 9. It is somewhat unusual in that the bandwidth is defined in terms of the points 3 dB up from the bottom of the notch. This is independent of the potentiometer setting, which controls the depth of the notch; a deeper null may be had at the expense of the notch being wider at the top. (Usually the width of the notch at the points 3 dB down from the top is more important; if so, the circuits given in Reference 9 are preferable.) Mohan[12] gives a variation of this circuit where the resistors and capacitors are equal and the center frequency may also be adjusted. He fails to point out that for his circuit gain is not the same above and below the notch, which is usually unacceptable.

A very general filter is the state-variable filter, also called by several other names in slightly different forms. One form is shown in Figure 10. This realization requires five op-amps, but gives low-pass, high-pass, bandpass and band-reject outputs simultaneously! Also, the frequency and damping may be adjusted as desired by varying resistors only. (In S-plane terminology, the poles may be placed anywhere in the left-half plane.) Thus, this filter could replace any of those described so far! However, the other filters have advantages, too, notably simplicity.

Several comments should be made concerning this filter. First, it should be obvious that many of the values may be changed without changing the general properties of the filter. The values shown are chosen to make the equations as simple as possible. For instance, if only the bandpass output is used and a gain of 1 is desired at center frequency, the input resistor may be changed from R to R_Q. Remembering the "virtual node" property, the input resistor clearly affects only the gain and not the frequency or Q.

Q for the band-reject is taken for the points 3 dB <u>down</u> from either passband. In the high-pass and low-pass, critical <u>frequency</u> damping means the highest Q for which there is no peaking in the frequency response curve, and it is the same for both high-pass and low-pass. This should not be confused with critical <u>time</u> damping which is the highest Q for which the step response of the low-pass filter has no overshoot. These are not the same value of Q, as indicated by the note.

The question is often asked, "How high in frequency can these active filters be used?" The answer, of course, depends on the amount of time,

talent, and money one is willing to spend, but the following is a good rule of thumb--at 1 kHz most anything will work. With care in selecting an op-amp and circuit impedances, most anything can be made to work up to 20 kHz. To get up to 1 MHz, considerable precautions must be taken, and it is preferable to use a circuit where the op-amp is in a very simple feedback loop, e.g., buffer, follower, etc. Above 1 MHz requires extreme care in virtually every aspect. These figures refer, of course, to the highest input frequency. For a state variable filter using ordinary op-amps and having a frequency of 20 kHz, the low-pass output may be fine but the high-pass output will give up so soon above 20 kHz that it will be closer to a bandpass. Likewise, remember harmonics. Do not expect to put a 30 kHz square wave into said filter and find anything like a square wave at the output, because that would require response near 1 MHz for the harmonics.

PASSIVE FILTERS

So, what does one do if the frequency is too high for an active filter? Fortunately, at high frequencies the size of inductors is reasonable so passive filters can take over. Theoretically, realizable filter responses are the same for passive and active. Passive filters normally must be terminated in a specific impedance at both input and output. Op-amps are still useful here as buffers for the passive circuitry.

The circuits of Figure 11 are handy. The low-pass version (Figure 11a) is sometimes called a "delay line" filter, as it is a lumped-parameter approximation to a true transmission line. Indeed, some of the same formulas are applicable. The important difference is that the lumped-parameter line will not pass frequencies above a certain cutoff, as is obvious from the series-L, shunt-C ladder construction. The circuit is most easily analyzed by breaking it up into symmetrical "L" half-sections as shown. The circuit has quasi-constant impedance; each section doesn't care whether it is attached to another section or to a terminating resistor. The characteristic impedance and delay per section are given by the same formulas as for a transmission line. The cutoff frequency is apparent from the L and C of the half-section.

The dual circuit is the high-pass filter of Figure 11b. The formulas are the same, except delay per section is not applicable. The dual of delay would be advance or prediction, which does not exist in the real world. The phase shift is inversely proportional to frequency, which apparently has no particular physical significance. These filters are discussed in more detail in Reference 13. Response curves are given. Also, active filter versions for lower frequencies are derived; these are also given in Reference 9.

OSCILLATORS

Since op-amps possess high gain, they are naturals for building oscillators. Again, a very large number of types have been built and the

general subject of oscillator design can become quite involved, so only a few examples are shown. The first example is a simple function generator (Figure 12a) consisting of a hysteresis clipper and an integrator in a feedback loop. The clipper output is, of course, a square wave switching rail-to-rail, or as far as the op-amp (or comparator) will go. Integrating the square wave gives a triangular wave ranging between the trip points of the hysteresis clipper. Clipping the triangle wave gives a square wave, completing the loop. Frequency does depend on the amplitude selected for the triangle wave, and it is 1/2RC for a peak amplitude of half the supply voltage, which case results for the values shown. Frequency and amplitude are reasonably stable, and oscillation is self-starting with no settling time required. (Here is an example where dc feedback is missing in the integrator but present around the entire loop.) If a sine wave is also required, it is usually obtained by passing the triangle wave through a nonlinear wave-shaping network employing diodes to round the peaks. Distortion (harmonics) is fairly high, and amplitude stability of the triangle wave is critical.

An alternate wave-shaping is the function generator of Figure 12b, consisting of an inverting clipper and two integrators in a feedback loop. The clipper output is again a square wave. Integrating once gives the triangular wave; integrating again gives a waveform composed of parabolic half sections. The latter is a fair approximation to a sine wave, the equivalent harmonic distortion being about 5%. This waveform is fed back to the clipper, completing the loop. The gain of the second integrator is made larger than the first to keep the two outputs at approximately the same amplitude. In the simple model, the amplitude is stable, once oscillation is started by an external stimulus but, in real life, it will either decay to zero or increase until saturation occurs. The oscillator can be preset and then left to run for a few cycles, or else continuous stability can be ensured by degrading the integrators to a finite dc gain and adding hysteresis to the clipper to sustain oscillations. The frequency of oscillation is dependent on the amplitude chosen, but it is on the order of $1/2\pi RC$.

A similar oscillator utilizing a linear inverter rather than a clipper is shown in Figure 12c. The outputs of the three stages are sine waves of equal amplitude, spaced 90 degrees apart in phase. The condition necessary for stable linear oscillation is that at some frequency the loop gain be unity and the phase shift be multiple of 360 degrees. In other words, if the loop is broken at any point, the output must be exactly equal to the input so the oscillation sustains itself. Of course, in real life, these conditions cannot be met perfectly, so the same stability precautions as above must be observed. If limiting is employed to stabilize the amplitude, it would normally be done at the output of the inverter. The output would be taken from the second integrator and would be relatively free from harmonics because of the double integration. Oscillator frequency is given by the same formula. This oscillator may be seen to be a special case of the state-variable filter. This paradox may be explained in terms of complex analysis in the S-plane (real and imaginary frequency axes). A useful way of analyzing filters is the placement of the "poles" and "zeros" (roots of the denominator and numerator polynomials, respectively) in the S-plane. The state-variable, low-pass filter as shown (and, incidentally, all low-pass

filters discussed here) is an all-pole filter, i.e., the low-pass function has no zeros (except at infinity), only poles (2 in this case) which all must be in the left-half plane. If all "damping" is removed from the loop, the poles are on the $j\omega$ axis and the filter has infinite Q. This is not a very useful filter; in fact, it is an oscillator. Any stimulus will cause ringing which goes on forever (in the ideal case).

Another form of this oscillator is shown in Figure 12d. Three identical degraded integrators are placed in a loop. This circuit gives three sine wave outputs spaced 120 degrees apart (normal three-phase spacing). The component values have been chosen so that the frequency is given by the same formula. Again, some precaution must be added to ensure stable amplitude.

Crystal-stabilized oscillators have become very handy with the advent of subminiature crystals and integrated circuitry. A simple op-amp crystal oscillator circuit is shown in Figure 13a. At its resonant frequency, the crystal appears as a low impedance and the positive feedback causes oscillations. The crystal may be represented as a series LC with some unwanted series resistance (actually a parallel group of several, representing the fundamental and all harmonics). The RLC in conjunction with the load is a very narrow bandpass filter, which has maximum transmission and zero phase shift at the center frequency. The circuit would oscillate at this frequency, were the op-amp perfect which it is not, usually having phase lag. The loop conditions are then met below center frequency somewhat, so most circuits oscillate at slightly less than the crystal frequency. This can be compensated by adding phase lead. In fact, phase adjustment can be used to vary the frequency slightly in either direction. The dc negative feedback path, which has no effect on oscillation if the time constant is much, much longer than the oscillator period, is included to bias the op-amp in the linear range when power is first turned on, assuring that the op-amp cannot remain in a saturated state, preventing oscillation. Load resistance must be chosen carefully according to the crystal used, or the circuit may oscillate at a harmonic of the crystal frequency or a frequency determined by stray capacitance or not at all. The circuit may be made more reliable by adding a resistor in series with the capacitor to limit the high frequency gain just enough to give oscillation with the crystal in place. If oscillations cease when the crystal is removed, it is a good indication that the circuit will not oscillate at an extraneous frequency. Older style crystals are usually above 1 MHz which is too fast for most op-amps, but the newer tuning-fork style crystals are available down to 10 kHz, easily in the op-amp range. It should be noted that the output waveform is a square wave; a comparator may be used.

If a sine wave is required, the signal may be picked off at the crystal output with a high-impedance amplifier. Otherwise, the circuit of Figure 13b may be used. An LC low-pass with a cutoff frequency about twice the crystal frequency is added to attenuate the harmonics of the square wave to give a nearly sinusoidal wave. This circuit also prevents oscillation at crystal harmonics. The circuit is uncritical if the Q of the RLC is near unity. A "cleaner" sine wave may be obtained with a high-Q LC tuned to the crystal frequency, but the tuning becomes critical and LC drift will result in large amplitude changes or complete failure.

Sometimes the frequency stabilizing element must be an LC resonant circuit (e.g., one of the elements varies to control the frequency). Figure 13c gives an LC oscillator. R_B and C_B give a biasing path as before. R_f gives positive feedback, with L and C acting as a tuned filter to limit the feedback to the resonant frequency. Variable inductors are available in the form of transformers where the inductance in one winding varies according to the current in a second control winding because the core partially saturates. Variable capacitors are available as back-biased semiconductor diodes; use of these requires the addition of a coupling capacitor to allow a dc voltage across the variable capacitor. In either case it is desirable to keep the amplitude across the nonlinear tuning element small. The circuit shown allows this since both tuning elements are across the op-amp (comparator) inputs.

REGULATORS

Sometimes a stable dc voltage lower than the available supply voltage is required. The simplest method of obtaining such is a resistor and zener diode. This voltage may vary considerably if the load current varies because the current through the zener diode changes and a zener does not have zero impedance. This may be avoided by inserting an op-amp in a manner similar to a follower, as in Figure 14a. The changing load current is now handled by transistor T_1. The zener serves only as a reference voltage for the op-amp. Transistor T_2 is necessary only if reverse load current can occur. If load current is small enough for the op-amp to handle, both transistors are unnecessary. (The double emitter-follower can be added to most op-amp circuits if high current drive is required without causing significant problems, provided fast transistors are selected. The base-emitter drop is compensated automatically by including it in the feedback loop.)

If the supply voltage is not steady, but varies because of ac ripple or change in load current, the output still varies. The circuit may be improved somewhat by substituting a constant-current diode for the resistor or paralleling a capacitor across the zener. A more sophisticated approach is to "boot-strap" the reference voltage from the output, as shown in Figure 14b, since the output is presumably more stable than the supply voltage. This becomes a bit tricky, as a positive feedback loop is formed which can cause two problems. The circuit might turn on in such a way that the output and reference voltage are both negative instead of positive, which is unfortunately a stable situation. This usually will not happen if ground is the op-amp negative supply, as shown. If it does, it can be cured by a "dribble" resistor from the reference to the raw supply voltage. Also, oscillation or "motorboating" may occur. The cure here is usually placing capacitors at the appropriate points. Note that the output voltage must be larger than the zener voltage. Special integrated circuits are available expressly for voltage regulation and are normally preferable to these do-it-yourself circuits. Most have an amplifier inside similar to an op-amp and use the techniques described here.

IMPEDANCE-TRANSFORMING CIRCUITS

An interesting, quite different class of circuits is that which performs impedance manipulations. Instead of performing operations on signals (voltages), these circuits perform operations on element characteristics (impedances). The first example is the negative impedance converter shown in Figure 15a. The input impedance of this circuit is the <u>negative</u> of the load impedance Z_1. This means that if the load impedance is a resistor, the entire circuit looking in at the terminals indicated will appear to be a negative resistor. Operation may be explained as follows. If a voltage e_{in} is placed on the input, the voltage at the op-amp inverting input will be brought up to the same voltage by the negative feedback. The current required to do this is determined by Z_1. An equal current will flow around the positive feedback path, since the two feedback resistors are equal and have equal voltages across them. This current will flow into the driving source. Thus, the current required of the driving source will be equal in magnitude to that for a simple passive impedance Z_1, but will flow into the source rather than out of it. One-ohm resistors are shown for simplicity; it can easily be shown that the equation is the same for larger resistors as long as they are equal. One must remember that although the output of the op-amp is connected neither to the input or the load impedance, it must remain in its linear range for all possible inputs or the circuit will not function properly. This is termed the "internal swing" problem and must be considered in any circuit where there is an op-amp output which is not observed directly, for example multiple op-amp circuits such as that of Figure 8d. Both voltage swing and current swing must be accounted for. This circuit requires the load impedance and equivalent input impedance to be referenced to ground. It cannot be made into a floating negative impedance by connecting the two terminals shown as ground to something else. (The previous report[1] incorrectly stated that it could.) The problem is that the current would flow into (or out of) both terminals at the same time; the circuit would not appear as a simple two-terminal device. A true floating impedance would require the addition of a second op-amp.

If the load impedance is placed in the negative feedback path (Figure 15b), a new sort of function results, that of the negative gyrator. The impedance at the input terminals appears as the negative <u>reciprocal</u> of the load impedance. The input <u>voltage</u> now causes a proportional <u>current</u> through Z_2. The <u>voltage</u> induced across Z_2 causes a proportional <u>current</u> into the driving source. Thus, the relationships of voltage and current are interchanged between the load impedance and the input. If the resistors have a value R instead of unity, the impedance is scaled by R^2, i.e., $Z_{in} = -R^2/Z_2$.

The effect of a positive gyrator is easier to understand, so let us design one by combining Figures 15a and 15b into Figure 15c. A gyrator can do two things. A linear impedance, usually a capacitor, will be transformed into its complement, an inductor in this case. A nonlinear impedance, such as a diode with the characteristic indicated, will have its current and voltage interchanged, as shown. If the negative gyrator of Figure 15b is used, the impedance is also made negative if linear and moved to a different quadrant if nonlinear.

Note that neither Figure 17b nor 17c will actually work as shown if the unspecified impedance is a capacitor because there is no dc negative feedback for the associated op-amp. This is true of many of this class of circuit. After the circuit is designed, one should step back and look at "what's really happening" to see if there is a basic flaw. Often it can be fixed without seriously affecting performance, for example, by the addition of a large resistor across the capacitor in this case. In contrast, Figure 15a does work for creating a negative resistance. However, whether it is stable depends on the resistance of the circuit it is connected to. If the source resistance is greater than the load resistance, the circuit will not be dc stable. This is readily apparent because, in that case, the positive feedback will be greater than the negative feedback. For a synthetic inductor circuit that actually works, see Reference 9.

By now the reader should be wondering what wonderful thing happens if the load impedance is placed in the positive feedback path. Unfortunately, this is just equivalent to the first circuit. The general circuit and the resulting equation is shown in Figure 15d. The equations using resistances other than 1 ohm stated above can be found directly from this formula. Figure 15e gives an alternate version. It is surprising that the inverting and noninverting inputs can be blithely interchanged and the same (theoretical) function results. One might suspect that one works and one doesn't. That is correct, but which one works depends on the application. The difference is, basically, that Figure 15d is short-circuit stable (input connected to ground) while Figure 15e is open-circuit stable. This is another generality with this class of circuit; often there are complementary versions and the proper choice depends on the application. Also, these equations indicate the possibility of impedance magnification or reduction. Indeed, this can be done by combining Figures 15d or 15e with Figure 15a, but the following simpler circuits are possible.

Figure 16a shows an impedance magnifier circuit. The input impedance is proportional to Z_5, but is larger, i.e., it requires proportionally less current for a given input voltage. The circuit is somewhat similar to Figure 9; the principle is similar. The lower terminal of Z_5 partially "follows" the input, by a fraction less than 1. Thus, Z_5 sees but a fraction of the input voltage, and it draws but a fraction of the current than it would if its lower terminal were grounded. This circuit has a handy feature in that the voltage at the input can be "remotely" sensed without loading it by observing the voltage at the output of the first op-amp, which is the same because it acts as a follower.

A complementary circuit is the impedance reducer circuit shown in Figure 16b. Operation is similar except the voltage at the lower terminal of Z_6 moves in a sense opposite to that of the input, so extra current is required. The impedance reducer makes an inductance or resistance appear smaller and a capacitance appear larger, similar to the Miller effect. The impedance magnifier does the opposite and is analogous to the bootstrap circuit used for high input impedance. Again the op-amp output voltage and current drive capabilities must not be exceeded, which places practical limits on the magnification or reduction achievable. For example, Figure 18b cannot be used to make a small capacitor act like a large one for use in a power supply. Close inspection of any proposed circuit will reveal that it simply won't work. Note that in either circuit, behavior depends only on the

ratio of the two resistances, which may be combined into a potentiometer for easy adjustment if desired. The appropriate formulas are given.

This class of circuits is not often used, and it is presented here principally as an indication of what can be done. The most common use is in active filters, covered in Reference 9. There are many other circuits published which perform these functions, usually under the title Generalized Immittance Converter (GIC). There seems to be an infinity of ways of interconnecting op-amps. The watchword is simplicity. Every feedback path gives another possible mode of oscillation. Oscillations can usually be cured by adding capacitance, but almost always at the expense of high frequency performance, often to an unacceptable degree.

From these circuits can be seen a larger class in which the voltage and current are not sensed at the same terminals. This class may be termed transimpedance circuits, or voltage-current converters. Figure 17a shows a current-to-voltage conversion circuit or transresistance. The <u>voltage</u> out is proportional to the <u>current</u> in. It should be noted that the transfer function has the dimensions of a resistance; it is not a dimensionless ratio as with the circuits discussed first. The inverter or summer circuits may be derived from this circuit by adding a resistance(s) in series with the input. If the inversion cannot be tolerated, the input may be placed at the positive input with the resistance to ground and the op-amp connected as a follower. However, this circuit loses the desirable property of the input being a virtual ground. At this point the reader might wonder how a signal (or information) can be transmitted into a virtual dead short to ground. The answer is that we are used to thinking of signals as voltages. Duality says we could also use currents, in which case it is desirable to keep the input impedance as <u>low</u> as possible, not high. The physical explanation is that the input voltage cannot be exactly zero. Connecting the inverting input directly to ground obviously would cause loss of signal.

Figure 17b shows the analogous circuit, the more familiar transconductance. Here current out is proportional to voltage in, and the transfer function has the dimensions of inverse resistance or conductance. Again, the limitations of the op-amp must be remembered; it cannot supply a large current into a high impedance because the voltage developed will saturate the amplifier. Often one side of the load impedance is unavoidably connected to ground so this circuit cannot be used. In this case, the more complicated circuit of Figure 17c may be used. Current into the load is "sensed" by observing the voltage across a floating resistor using an instrumentation amplifier with one buffer omitted (because it does nothing). This voltage and, hence, the current, is made proportional to the input voltage by feeding it back to the input amp. Feedback is indeed negative even though the noninverting terminal is used for feedback. Likewise, coming directly into the inverting terminal of the input op-amp with no apparent feedback is acceptable. The circuit would appear more conventional if the inputs to both the input amp and difference amp were interchanged.

Figure 17d shows a single op-amp circuit invented by Howland which also does this. Both inputs are used, so the circuit may be made inverting or noninverting by grounding the unused input, or differential. Note that the circuit relies on matching of the resistor pairs for proper operation. Also, it relies on the load impedance for stability; without it, negative and

positive feedback paths are equal. Note that the use of ground as a second input, usually of opposite polarity, is possible in many circuits, for example, Figure 17b.

RECTIFIERS

Between linear and nonlinear circuits is a class termed piecewise-linear. Probably the most important type is rectifiers, which are frequently used in signal processing systems. A large number of circuits have been published; indeed one author (J. Graeme) has over a dozen. All are variations on the same basic principle; only a few will be shown here.

Figure 18a shows a noninverting, half-wave rectifier circuit. For positive inputs the diode conducts and the circuit acts as one big follower. For negative inputs the diode opens and the output remains at ground. By placing the diode in the feedback loop, its forward drop is effectively reduced by the gain of the op-amp, making it negligible. The discrimination between positive and negative is done by the op-amp input rather than the diode, so offset errors are on the order of millivolts rather than hundreds of millivolts. To detect negative inputs, the diode is reversed. The maximum differential input voltage of the op-amp must not be exceeded. In these circuits the op-amp really must also perform as a comparator, a very demanding application.

An inverting rectifier is shown in Figure 18b. It is generally preferable for several reasons. It requires only one op-amp. However, when the diode is not conducting, the op-amp is out of that loop and the output impedance is equal to the feedback resistor; another op-amp may be necessary for a buffer anyway. The gain may be made greater than one by changing the resistors. The extra diode prevents the op-amp output from slewing all the way to the positive supply voltage when the rectifying diode is not conducting; this would cause a long recovery time back to the linear mode due to the slew rate limit of the op-amp, which is a serious problem in rectifier circuits. Differential input range is of no importance here. Again, an output of the opposite polarity may be obtained by reversing both diodes.

A full-wave rectifier can be made by adding the input to the output of a gain-of-two, inverting, half-wave rectifier, using a summing inverter. A simpler circuit may be made by removing the clamping diode from the inverting half-wave, as shown in Figure 18c; the input simply goes straight through when the diode is not conducting. However, this circuit suffers from all three problems: op-amp input restriction, an impedance of both resistors in series half the time; and the slew rate limitation.

The dynamic range of these circuits is limited on the low end to a few millivolts by the op-amp input offset. Belanger[14] has pointed out that this may usually be avoided by ac coupling the op-amp. Figure 18d shows a simplified version of his circuit. The resistor and capacitor connected to the inverting input are not labeled since they are generally made as large as possible. Offset can be further improved by adding a resistor at the non-inverting input equal to that at the inverting. The op-amp together with

these components can be thought of as an ac op-amp, as indicated by the dotted line. The circuit is not a true full-wave as claimed, as the output still requires a differential amplifier. It does, however, point out that both halves of the waveform can be available simultaneously.

DIGITAL CIRCUITS

Sometimes in small mixed systems it is desirable to perform some digital logic without using regular digital circuits because of supply, interface, parts count, etc., problems. For example, an analog system may be required to change its mode of operation on command. If the command is pulse rather than level, memory is required. There may be a leftover op-amp; it can provide this function. If necessary, op-amps can perform many simple digital functions, as shown in Figure 19. In Figure 19a the positive feedback makes the op-amp latch to act as a flip-flop, which may be set or reset by inputs which override the feedback. Here all inputs and outputs swing between +V and -V, the supply voltages. A number of variations are possible. This circuit could be modified to be set or reset by positive or negative pulses, respectively, on a single control line, which is not possible with conventional digital circuits. These circuits can use a single power supply, as is usually the case with digital logic. The rule of thumb is that any resistor to ground is split into two of twice the value, one to ground, and the other to supply voltage. This biases all inputs at half the supply voltage but leaves operation unchanged otherwise.

Figure 19b shows a pulse generator, used where the output pulse required is shorter than the input pulse. Figure 19c shows a delay multivibrator, which is used in the opposite situation where the output pulse must be longer than the input pulse. Figure 19d shows an astable multivibrator. Note that this could also be considered a square wave oscillator. This circuit has several advantages over the conventional 2-transistor multivibrator: it is self-starting, the output is a clean square wave, and the frequency is not affected by supply voltage or loading. Formulas for the periods are for rough calculation only. They use the approximation $1/e \approx 1/3$ and assume the op-amp output swing is rail-to-rail. Note that these circuits are strictly nonlinear, as are the ones of the following section.

LOGARITHMIC CIRCUITS

The usefulness of logarithms is well known: the ease of handling large dynamic ranges, the ability to perform multiplication, division, root extraction, raising to a power, root-mean-square, etc. An electronic circuit which can perform log or antilog conversion is often useful in signal processing systems.

An ordinary silicon semiconductor diode has a logarithmic current-voltage characteristic over a fairly wide range. However, most circuits have voltage inputs and outputs and diodes are passive devices and work at rather

low voltages, so op-amps find another use here. A diode (e.g., 1N645) has approximately the characteristic V = 1/10 log I (V in volts, I in nanoamps, logarithm to base 10). Thus, the circuit in Figure 20a has the relationship V_{out} = log V_{in} (V_{out} in volts, V_{in} in microvolts, logarithm to base 10), and the circuit in Figure 20b has the relationship V_{out} = antilog_{10} V_{in} (V_{out} in microvolts, V_{in} in volts).

Thus, these two circuits in combination with summing amplifiers can theoretically perform any of the operations mentioned above. However, the practical problems, such as temperature variation of the diode characteristic, have not been considered here and these are not acceptable as shown. Special IC's are available for log/antilog conversion and also for multiplying, dividing, squaring, square rooting, and RMS, and these would normally be easier and better to use. They usually use the basic principle shown here.

Several problems, common to all logarithmic circuits, should be mentioned. Because of the wide dynamic range on the log end of the circuit, units to be used are a problem. Hence, the use of microvolts in these examples. Since the circuits are nonlinear, error must be carefully thought out. For example, a dc offset at one point does not necessarily translate to a corresponding dc offset later on, but possibly a gain error. Logarithms are not defined for negative numbers. What is usually done circuit-wise is to restrict the range (which is a practical necessity anyway) and use positive outputs to signify positive inputs, and vice versa. This may be done for the circuits shown by paralleling each diode with another in the opposite direction. In fact, as shown, Figure 20a works only for positive inputs and Figure 20b works only for negative inputs, so they are not compatible. Similarly, the logarithm of zero is not defined, so grounding the input of a log circuit will not necessarily produce a meaningful output.

SPECIFICATIONS

If an ideal op-amp could be built, only one type would be necessary and we would never need a "spec" (specification) sheet. However, none of the characteristics of an op-amp will be ideal, and most must be specified. Also, although the op-amp symbol used here has only three terminals, actual devices have five or more. The use of extra pins will be explained under the limitations to be discussed shortly.

There are five general rules of caution to observe when using the spec sheets. The first is to read the entire spec sheet, including all the footnotes. Definitions and testing methods may differ between manufacturers. Each type of op-amp has its own peculiarities. Also, make sure you have the spec sheet for the right device. A given type may be modified during production, although this is usually indicated by a change in part number. The second warning is to design on the basis of worst-case values if possible. The "typical" spec is no more than a "maybe," but the devices are supposed to always meet worst-case specs. The third rule is to observe the temperature specs. The devices are almost never used at the same temperature the manufacturer tested them. Fourth, pay no attention to general descrip-

tions such as "high-performance." There apparently has never been a "low-performance" or even a "medium-performance" op-amp. Last, don't use oddball devices unless absolutely necessary; some particular parameter may be fantastic, but often there are bad side effects that will cause grief.

Table 1 gives a summary of the common limitations of op-amps. These limitations will be presented in more detail here. No attempt is made to define the parameters precisely as some definitions are not standard and others depend on the test conditions. Refer to the appropriate spec sheet for definitions.

INPUT IMPEDANCE consists of two parts, although frequently only one is specified. The DIFFERENTIAL input impedance is effectively an impedance across the two op-amp inputs. The number given may not seem very high, but this is not of serious consequence in linear circuits because no appreciable voltage is developed across the inputs. In comparator circuits it may be a problem. The COMMON-MODE input impedance is effectively an impedance from each input to ground. It is usually much higher than the differential and is often not specified. If only one number is given, it should be the differential. However, the common-mode impedance is usually not improved by the feedback.

INPUT CURRENT runs from a few microamps to almost zero. It is always there, so a dc path must be provided for each input.

OFFSET CURRENT is the difference in current required between the two inputs. This difference must be less than the worst individual current, and it is usually considerably less since the input transistors should be well matched.

OFFSET VOLTAGE similarly is an inherent voltage difference, typically around a millivolt, between the two input transistors. Some op-amps have extra terminals for attaching a potentiometer to balance out offset. However, temperature effects will still be present and, in fact, may be worse as the balance control may compensate for the input by unbalancing another part of the circuit.

INPUT VOLTAGE RANGE is the maximum voltage that may be applied to an input without misoperation or damage. It is not a problem with inverter type circuits where the input is a virtual ground, but it may be with follower type circuits where the inputs have the same dynamic range as the output. Newer designs usually allow a swing nearly equal to the supply voltages. However, one must distinguish between: the range over which the op-amp will not be damaged, usually given under "maximum ratings" and usually equal to the supply voltage; and the range over which it will work properly, which is usually given under "electrical parameters" and is usually less than the supply voltage. The worst manifestation of the latter is that if the output range of the op-amp is greater than its input range (which is usually the case) and exceeding the input range causes some internal inverting stage to saturate and become noninverting (sometimes the case), the op-amp inputs effectively become interchanged and the circuit may latch up. Note that this may be induced not only by signals but noise transients picked up from other circuitry (including machinery, power lines, etc.). The cure is to limit the

op-amp output swing, or at least the fraction of it fed back to the input, or switch to another type op-amp.

DIFFERENTIAL INPUT VOLTAGE RANGE similarly is the maximum allowable voltage difference between the inputs. This is a problem only with the nonlinear circuits, as in the linear circuits the two inputs are at nearly the same voltage. Some op-amp types have back-to-back diodes across the input to protect the transistors; these have a current limit rather than a voltage limit. Many newer designs allow full supply voltage range.

OUTPUT CURRENT DRIVE determines the minimum impedance that may be driven as a load without faulty operation. In a few older devices, excessive loading may also cause damage. It must be remembered that feedback components may also appear as a load to the op-amp. Most op-amps will supply several milliamps in either direction, but this may be significantly less for low-power op-amps.

OUTPUT RESISTANCE is typically on the order of a hundred ohms. This figure sounds high but, in a linear circuit, it is reduced by the negative feedback so the dynamic output impedance of a typical circuit is usually less than an ohm. Remember, however, that this applies only when the limitations of the op-amp are not exceeded and it is still in its linear range. It has no direct connection with maximum output current. An instance where output impedance is important is capacitive loading. The equivalent RC lag can cause the circuit to oscillate; this will be covered later under "compensation."

OUTPUT VOLTAGE SWING depends not only on the loading but also on frequency, as will be explained later. For reasonable loads at low frequency, the output swing is nearly equal to the supply voltage for most designs; sometimes it is equal.

COMMON MODE REJECTION RATIO may be thought of as the ratio of differential gain to common mode gain (both inputs tied together). This ratio runs 60 dB to 100 dB, so the assumption that the op-amp is insensitive to signals common to both inputs is valid in all but the most critical differencing applications.

SUPPLY VOLTAGE REJECTION RATIO implicates some more terminals which are usually ignored and often do not appear in the diagrams--the supply voltages. Being a power gain device, an op-amp requires supply power. This normally takes the form of one positive and one negative dc voltage of equal magnitude: $\pm$15V is standard; $\pm$12V is frequently used; some op-amps will operate as low as 1V total. (A few require power supply ground.) In the real world, this dc is not completely constant but has hum, ripple, and noise. The power supply rejection ratio is the ratio of the effect for a given (differential) change at the input divided by the effect for the same change on a supply voltage. This ratio for low frequency (drift, hum, ripple) is similarly 60 dB to 100 dB, so this problem can be ignored except in low-signal, low-noise applications, typically high-gain (pre) amplifiers. Another problem here is that the later stages impress a small amount of signal on the (imperfect) power supplies. Part of that is fed back to the earlier stages through the power supplies. If the gain in between is high enough, the system oscillates. Cures are: more capacitance on supply leads (best at

high frequency), or decoupling with regulators (best at low frequency), or a combination.

VOLTAGE GAIN is typically 10,000-1,000,000 (80 dB - 120 dB) so it is usually not a problem, with two notable exceptions. The first is when the op-amp is used in a high gain amplifier stage. The second limitation is the following.

FREQUENCY RESPONSE is the gain of the op-amp as a function of frequency, and it is usually given as a series of graphs. The first graph is the gain of the op-amp by itself, termed the open-loop gain. A typical graph of an op-amp before compensation is shown in Figure 21a. The response is uniform (flat) at low frequencies where no frequency effects occur. Past some breakpoint frequency, f_1, it drops off at 6 dB/octave, caused by one internal state acting as a low pass. Past a second point, f_2, it drops 12 dB/octave where a second stage does likewise. These breakpoints continue until all stages are accounted for or gain is less than unity (0 dB). When making calculations, the op-amp gain _must_ be taken at the frequency of interest. Phase response is usually also shown. Additional curves of closed-loop gains may be shown. Internally compensated op-amps have only one open-loop curve, typically like the dotted line in Figure 21b. Externally compensated ones will have several which depend on the compensation used. This will be discussed in more detail later.

SLEW RATE. If the input voltage is changed instantaneously, the output voltage will slew at a finite rate, usually in a linear fashion. The slope is called "slew rate" and places an important upper limit on high-speed performance. The op-amp may have gain up in the megahertz region but still be useless because the output swing is limited. One volt/microsecond may not sound slow, but it means the normal $\pm$10V output swing cannot be obtained on a 20 kHz sine wave. Sometimes the specification is for power bandwidth, the highest frequency for which the op-amp will produce a full-amplitude sine wave, which bears a direct relation to slew rate. For externally compensated op-amps, slew rate is usually a function of compensation. Fast op-amps usually have both good frequency response and good slew rate, but the two are not directly related.

BANDWIDTH (small-signal) is usually specified as the frequency that the gain of the op-amp falls to unity (0 dB). Of course, the op-amp is useless at that point, but it is an indication of high-frequency performance. Furthermore, since modern designs usually exhibit a 6 dB/octave (20 dB/decade) slope over most of the range (constant gain-bandwidth), the gain at other frequencies can be readily calculated.

SETTLING TIME is the time for the op-amp, in a given circuit, to reacquire the "correct" output value after a sudden change in the dc input, i.e., a step. This is specified in many different ways, often so as to make a particular device appear attractive. Take it only as a rough indication of how it will work in your particular application. Alternately, rise time may be specified. Either depends on several factors and cannot be readily calculated.

INPUT NOISE. Originally op-amp noise was not specified at all; it was simply conceded that op-amps were not good. Later it was specified sporad-

ically, in a number of ways, usually not too helpful. Manufacturers have pretty much now standardized on a good, simple way. The equivalent input noise spectrum is specified both for voltage and current. Total noise is found by integrating across the portion of the frequency spectrum used; typical operation is in the flat portion of the spectrum, so this amounts to multiplying the voltage level given by the square root of the bandwidth. The noise current is multiplied by the impedance at each input and these, together with the noise voltage, are added root-sum-square (orthogonally). Like offsets, these parameters are referred to the input of the cp-amp. Indeed, they are handled much the same way; they are the ac parameters that correspond to the dc offsets.

Another source of noise is "popcorn" noise, so called because when amplified and played through a speaker it sounds like popcorn popping. It can be described as random sudden changes in dc level; it is a transient phenomenon. It is believed to be associated with surface defects; manufacturers are learning how to avoid it, and it should not be a problem in the future. It is difficult to specify and is usually ignored.

SUPPLY CURRENT. The designer of a circuit usually needs an estimate of the current drawn in order to specify the power supply, and possibly the heat dissipation. For an op-amp the supply current is specified; it will be specified at a given supply voltage, but usually remains fairly constant for the range of supply voltage over which an op-amp will operate. Power dissipation is, therefore, the supply current multiplied by the total supply voltage. Current is the same for both supplies, as most op-amps do not use ground directly. This does not include the load, which is normally referenced to ground. Modern op-amps require so little "quiescent" current that the load current may be comparable to the op-amp current; note that load current depends on the output signal. For quad op-amps the total package current is usually specified, but the current per op-amp is sometimes given because that figure looks better. For op-amps where the supply current is "programmable," it is given by a formula or graph. Op-amps are now available which will operate at very low supply current, but speed goes down virtually in proportion to current.

TEMPERATURE COEFFICIENTS. An integrated circuit seldom operates at room temperature; even in a controlled environment there is usually some self-heating from power dissipation. All parameters are functions of temperature; fortunately the variation is slight except for a few, and those are listed. The most serious is that the input balance, voltage and current offset, will change with temperature. An external balance circuit can null the offset only at one temperature and, in fact, may make the temperature coefficient worse. Alternately, the parameters may simply be specified for worst-case over the entire temperature range.

OTHER MAXIMUM RATINGS. There are other maximum ratings such as supply voltage, power dissipation, and operating temperature which must not be exceeded; but these seldom pose a limitation.

INTERNAL CIRCUIT DIAGRAM. Although an internal schematic is supposedly not necessary, this author recommends against using a device for which the manufacturer does not give one, at least a simplified version. Not only does lack of a diagram make it near impossible to figure out what is happening if

the device misbehaves, but it seems the devices without diagrams invariably are the ones most likely to do so.

PRICE. This important specification is omitted from nearly all spec sheets, partly because it is the one parameter subje to considerable change. There can be quite a difference between new, exotic types and the older "jellybeans," but most monolithics are low enough that cost difference between types is not terribly significant. The cost difference between grades of one device, i.e., commercial to full military, can be greater.

COMPENSATION

Frequency compensation is a modification of the frequency response curve necessary to make the op-amp stable in the negative feedback (closed-loop) configuration required for all linear circuits. There are three general approaches to the problem. The simplest is to use an op-amp that is internally compensated by the manufacturer. His open-loop response curve will include this compensation and, if response is adequate, use it. However, this compensation is necessarily designed for worst-case conditions and, hence, this type of op-amp is slowest and suitable for the smallest number of applications. The second approach is to find a manufacturer's circuit close to that desired and use the suggested compensation. This nearly always works properly and covers quite a variety of applications. The manufacturer will give either the resulting closed-loop response curves or the open-loop response curves for the op-amp with the compensation added, from which the closed-loop curves may be deduced. If all else fails, one is forced into the last resort--understanding the problem and designing suitable compensation. To that end some basic compensation techniques are presented here.

The compensation problem arises as follows. Referring to Figure 21a, at low frequencies the amplifier only amplifies and the negative feedback loop has 180 degrees total phase shift around it as intended. Past f_1 the amplifier introduces 90-degree phase lag due to an internal stage acting as a lowpass, but total phase shift is only 270 degrees and the loop is still stable. However, past f_2 another 90 degrees brings the total to 360 degrees. If the loop gain is still greater than one ($f_3 > f_2$), we have an oscillator rather than whatever we planned on. In fact, if the gain is only slightly less than one at the 360-degree point, although the circuit will not oscillate, it will still be highly underdamped and will ring badly on transients. The difference between the actual gain and 0 dB at the 360-degree point is referred to as the "gain margin"; the greater the margin, the more stable the op-amp. Alternately, the "phase margin" may be specified: the difference between the op-amp phase shift and 180 degrees at the 0 dB gain point.

Thus, for a follower (the worst case) the 180 degrees, 0 dB point on the op-amp frequency response is a nasty point, to be avoided at all costs. The simplest, "brute-force" method of doing this is to place a large capacitance at some point in the op-amp circuit which changes the response as shown in Figure 21b.

This is termed "lag" compensation and is used in most internally-compensated op-amps. This method maintains the 6 dB/octave slope past the 0 dB point at great expense of op-amp speed; hence, the limited application. As a practical consideration, the capacitor should be placed near the op-amp input where the voltage swings are small so that high currents are not required to drive the capacitor, limiting slew rate.

A more sophisticated method is "lead-lag" compensation which consists of a series RC instead of simply a C. This reduces the gain at a 6 dB/octave rate until it is below the 0 dB point, then has no further effect. The new curve is shown in Figure 21c. If the capacitor is omitted, the op-amp is "compensated" by simply lowering the gain at all frequencies. This is acceptable in some applications, and it is often suggested for "decompensated" types.

A third method is "lead" compensation which maintains the 6 dB/octave slope to beyond the 0 dB point by increasing the high frequency gain of one stage, which adds phase lead. This is shown in Figure 21d. This method has the nice property of "speeding up" the amplifier. However, it can be seen from the graph that it is rather tricky, as it extends the response to higher frequencies where additional breakpoints may occur anyway. It is not applicable to all op-amps, whereas the first two methods are. A related method is "feed forward." For a stage (or group of stages) which is noninverting, above the frequency where the gain falls below unity, it is advantageous to simply bypass the stage. This eliminates a phase lag, which eases compensation. Again, this is a rare case of having your cake and eating it too, and it is not applicable to all op-amps.

Some texts point out that an op-amp may be modeled as an integrator. This is actually a better model than a high-gain amplifier, especially for internally-compensated op-amps (refer to Figure 21b). However, it is more difficult conceptually, so generally not preferred. It does point up why the differentiator configuration is basically unstable: the op-amp and the feedback network can both be represented as integrators, and they are in a loop. It also shows why capacitive loading frequently causes oscillations: the shunt load, in conjunction with the effective series resistance of the output impedance, gives a lag network within the loop. The same is true of op-amp input capacitance in conjunction with the feedback resistance. Fortunately, both tend to kill the loop gain at high frequency, so usually the loop gain is less than unity at the frequency where 180-degree phase shift accumulates.

The above analysis was for the follower circuit where the feedback ratio is unity and the loop gain equals the op-amp gain. For an amplifier stage gain of 10 (20 dB), the feedback ratio is about 1/10 (-20 dB). This may be taken into account by moving the 0 dB line on the op-amp response curve up 20 dB, and so forth, for other gains. Thus, the higher the stage gain, the lower the loop gain and the easier the compensation. Hence, for high stage gains the internally compensated op-amp is a bad choice, as indeed compensation may not be necessary at all!

EXAMPLES

Now let us see how these restrictions (specifications) affect some practical circuits. Consider the simple clipper of Figure 22a. Usually any dc level on the signal is unwanted, so a capacitor ac couples the input. A resistor to ground is then necessary to provide the input current. This gives an offset of (R_+) (i_+). This may be minimized by placing an equal resistance at the other input. The offset then becomes R $(i_+ - i_-)$ which is normally significantly smaller. If the resistors are not perfectly matched, there is another error term equal to the difference in resistance times the (average) input current. In recent designs the two input currents are so well matched that this can become the predominant term.

This, of course, may be minimized by using an op-amp (or comparator) having a small input current. However, one is still faced with the input voltage offset, which gets added to the offset caused by current. Furthermore, older low-current designs like a Field-Effect Transistor (FET) or Darlington input op-amps will have a worse voltage offset than normal. These offsets typically amount to several millivolts, so the signal will not be clipped well unless it is much greater than this. However, there are new designs which excel in both parameters.

As pointed out, the clipper is about the simplest possible application for an op-amp. However, the optimist who thinks this means nothing can go wrong is sadly mistaken. Consider the following additional problems. An ideal clipper gives a square wave output for a sine wave input. A perfect square wave does not exist in the real world because the frequency spectrum would be infinite. So what really comes out? First, the square wave will have finite rise and fall times, as in the vicinity of the input zero crossings the op-amp is only amplifying a finite slope with a finite gain. For a 1 millivolt peak sine wave at 1 kHz input, a ±12V output swing and an op-amp gain of 100,000 (100 dB), the output will have a rise time of about 20 microseconds, which may not be acceptable. An uncompensated op-amp (or comparator) should be used here. Even for large input voltages, the slew rate will limit the rise time. The rise and fall times probably are not equal, causing asymmetry which give rise to even-order harmonics which the square wave is not supposed to have. There is delay between the input zero crossings and the output zero crossing, again probably unequal. These inaccuracies are negligible in many applications, but one should consider the accuracy required before using just any circuit.

A more difficult problem which is virtually impossible to calculate in advance is this: at the zero crossing, the op-amp is momentarily in its linear range and exhibits a very high gain, possibly in the millions. If a millionth of the output gets back to the input through unwanted capacitive coupling or leakage, the circuit may oscillate or latch up. Necessary precautions may be more physical separation of input and output leads, special ground-plane shielding, reducing circuit impedance at the input, supercleaning PC boards, or interchanging the op-amp inputs, if possible.

Next consider the linear amplifier of Figure 22b and assume a gain much greater than one is desired ($R_f \gg R_{in}$). The feedback is ac coupled for the following reason: the input offset voltage is amplified exactly as an input signal. If the capacitor is removed, an input offset of 5 millivolts and a gain of 1000 would give an output offset of 5V, which is probably unacceptable. With ac coupling, the dc gain is unity and the output offset is only 5 millivolts (plus offset due to input current). This points out why high-gain dc amplifiers are difficult to build. If the driving source can supply dc current and its offset is small, the RC at the noninverting input may be unnecessary.

There is also a practical limit on the amount of ac gain that may be obtained. It is normally desired that the gain be determined by the resistors, i.e., be unaffected by op-amp gain which is highly variable. This can be assured by making the gain of the stage much less than the guaranteed minimum gain of the op-amp. The overall gain change is reduced approximately by the "excess gain" of the op-amp, equivalent to "loop gain." For a stage gain of 60 dB and an op-amp gain of 80 dB, the effect is reduced 20 dB. A 50% change in op-amp gain will give roughly 5% change in stage gain, which may or may not be acceptable. Note that the 80-plus decibel op-amp gain is required at the frequency of operation, which judging from the circuit, is 1 kHz or above. (Below this frequency response starts to fall off, determined by the time constant of R_{in} and C.) An ordinary "garden variety" op-amp will not do.

Figure 22c gives a circuit which is quite useful and also involves several of the practical problems discussed so far. It is an Automatic Gain Control (AGC) amplifier. An AGC amplifier varies its gain automatically to maintain the output at a preset level if the input level changes. The circuit operates as follows. The top op-amp is set to a gain of 100 (40 dB). The negative peaks of the output are detected by the diode. If this average exceeds the threshold set by the resistive divider the bottom op-amp, which is an integrator, slews positive. This turns the FET on, creating a voltage divider at the input which brings the amplitude back down to the preset level (about 1 VRMS output for sine wave). The reverse occurs if the signal amplitude drops.

Now for the practical limitations, starting with the range over which proper AGC action occurs. The minimum input signal is 10 millivolts RMS, determined by the gain of the fixed amplifier. This, of course, can be changed by changing the feedback, limited again by the op-amp gain. The maximum input signal is determined by the range of FET. The "on" resistance of this particular FET is about 10 Ω, giving a range of 1000 (60 dB) for a maximum input of 10 VRMS. This range may be increased by increasing the input resistor, but the capacitance of the FET may affect frequency response. For signals outside the range, AGC action ceases and the circuit operates as a fixed-gain amplifier.

One must also consider what happens when the AGC is driven outside its linear range. It is desirable that the integrator stop slewing. If it slews way past the linear range, it will take a long time to slew back, causing an unduly long recovery time when the input signal amplitude returns to AGC range. In this circuit, if the input amplitude is too large, the integrator tries to slew positive until the field effect gate becomes forward biased.

The Schottky diode prevents this. If the input is too low, the integrator will slew negative until the FET ceases conduction. If a normal op-amp requiring positive input current is used for the fixed amplifier, its output will go negative because of the loss of an input dc bias path. This will appear to the detector as a signal, stopping the integrator slew. Thus, the integrator stops precisely at each end of the useful range of the FET. The AGC time constant is about 20 milliseconds for low input amplitude; longer for high amplitude. (It is related to the integrator time constant but is affected by the FET characteristic so is not exactly equal.) It may be changed by changing the value of the capacitor in the integrator.

The amplifier is ac coupled, as any dc offset causes error in the amplitude detector section. A 741-type is used for the integrator, as worst-case compensation is necessary and high speed is not required. The capacitor to ground at the output prevents possible high frequencies being coupled from the input through the capacitance of the FET to the integrator output, as the slow op-amp cannot handle these. It has no effect on integrator time constant. A 741-type is not optimum for the amplifier as compensation may not be necessary due to the high gain of the stage. A 741-type is marginal here at 1 kHz, but gain error does not matter much as it is corrected by the loop anyway.

NOMENCLATURE

Each manufacturer seems to use a different system of identifying devices. (There are two exceptions, discussed at the end of this section.) The generic code or basic type is usually a three-or-four-digit number buried in the middle of a longer type number; some newer units have two (e.g., National) or five (e.g., Motorola) digits. Usually there is a letter prefix indicating the manufacturer, and often whether the circuit is analog or digital. Fairchild, the first manufacturer, began theirs with the Greek letter μ (for micro) which is often not available, especially with computers. It is often written out, mu, or abbreviated M, or left out altogether, further confusing the issue. The package type is often indicated in a suffix, usually letters but sometimes a mixture of letters and numbers. The grade (quality) of the chip is usually indicated in the suffix by letter, but sometimes by changing the prefix (e.g., Signetics), and sometimes by a change in the basic part number, usually the first of three digits (e.g., National) or the second of four (e.g., Motorola). Any additional information is usually in the suffix. Some manufacturers also assign a part number for each specific type which may or may not resemble the type number.

In addition, there should also be a standard lot code which identifies the particular batch coming from that manufacturer. It is a four-digit number with no prefix or suffix. The first two digits are the last two digits of the calendar year made; the last two are the lot, up to 99 per year. Bad parts are traced by lot--if many turn up from a given lot, the entire lot is discarded or at least checked.

The military has its own system, which has two levels. The first is MIL-STD-883 which basically defines manufacturing methods (cleanliness,

inspection, package seals, etc.). Any manufacturer may claim to apply this to any part, usually indicated by a -883 on the end of the part number. Thus, there are many of these available at a modest increase in price which presumably gets you better reliability. The individual manufacturer must be consulted for details. Some also add a reliability grade. The next level is MIL-M-38510. It specifies everything about the device, of which 883 processing is a part. The part number will not look anything like the equivalent commercial number. The number will contain 38510 followed by a slash followed by more letters and numbers which define the particular device, including reliability level (hence, the term "slash sheets"). The number guarantees interchangeability no matter who the manufacturer. However, there is no guarantee that anybody makes it or ever will. There are relatively few of these available in analog, and the prices tend to be high.

TYPES OF OP-AMPS

Many types of IC op-amps are available, each designed for a purpose. This is done chiefly because some parameters can be optimized at the expense of others. Higher maximum ratings can be obtained at reduced performance; increased power dissipation can give higher speed; cruddier op-amps are cheaper to build, etc. Two nontechnical parameters must also be considered: availability and price. Attempting to stock every kind of IC op-amp ever made gets out of hand quickly. Likewise, a circuit requiring a $100 op-amp may sometimes be modified with a few $.10 components to use a $1.00 op-amp. Also, any device is useless if you can't get one when you need it. An important consideration is whether the device has a second source or direct substitute; one does not want all one's eggs in the same basket.

Table 2 is a list of the types of monolithic op-amps presently used in the Acoustic Signal Processing Branch (U25) and their general characteristics. This is included as an example and is <u>not</u> intended as an evaluation of manufacturers of components. It is necessarily slanted toward the older types, and some are of historical interest only. Manufacturers left out are not necessarily undesirable. The question is frequently asked, "Which is the best op-amp available?" The answer is simple--there is no such animal; it depends on the application. As for which is best for a particular job, the best op-amp is the cheapest, simplest, most available one that will meet the requirements.

There is now a bewildering number of op-amp types on the market. More types have been introduced since 1980 than were available at that time. The various types may be fairly well grouped into five categories, however. The first is "obsolete," meaning primarily that the type would not be considered for inclusion in a new design. It may still be available and may work fine. Second is the "741-type." This would be a general-purpose op-amp, fully compensated, having around a megahertz bandwidth and a volt per microsecond slew rate. Third is the FET/fast-741/decompensated group. These are generally around one order of magnitude faster than a 741. The FET input types are inherently faster. They have virtually no input current or current noise at the expense of worse voltage offset and voltage noise. The fast-741 is simply an improved 741, inherently somewhat more inclined toward instability. The decompensated types are not compensated for unity gain,

which inherently makes them faster. They generally can be "fudged" to work at unity gain anyway. The fourth group is low-power. These are typically programmable and/or designed to work at low supply voltage. They are generally slower than a 741, but some are improved 741's which have the same speed at an order of magnitude less supply current. The last category is special-purpose, any type that does not fit into the other categories.

It is impractical to try to summarize the device parameters; the reader must refer to the spec sheets. Instead only the (original) manufacturer, general description, and comments are listed. The table contains a term which did not occur in the previous editions of this report, "discontinued." One problem with a maturing technology is a "shakeout," widespread discontinuing of devices (and sometimes companies) that are not productive enough. In fact, some devices were discontinued shortly after announcement. A brief discussion of most of the various types follows, which should point out some of the practical problems with op-amps.

The 702 was the first monolithic op-amp; it was hard to use and easily destroyed; the 712 was a slightly more rugged version. The 709 was next, the first approaching general-purpose. But it could latch up in a follower configuration and was only partially protected. "Protected" here means either input will tolerate any voltage between the normal supply voltages (±15V), and the output will tolerate an indefinite short to ground or a temporary short to either power supply (one op-amp at a time for a quad). Also, "obsolete" means only that this author would probably not use it in a new design; it may still be available and/or in use.

The 715 was the original high-speed. Note again that "high-speed" involves both bandwidth and slew rate, which are not directly related. A given application may require either or both, but a "high-speed" device may be lacking in one or the other. The 725 was the original having outstanding dc characteristics; use of the term "instrumentation" here is unfortunate and should not be confused with the three op-amp configuration. The 735 was the original low-power; bias was fixed. The 739/749/381 are low-noise designs intended for stereo preamps, but they can be used as op-amps. The 740 was the original FET input. A JFET input reduces the input current to near-negligible level at the expense of voltage offset; however, the 740 never received widespread acceptance.

The 741 was the coming-of-age of the monolithic op-amp; it established general acceptance of the technology. It was the first fully monolithic, fully protected, fully compensated design; even today any device having these qualifications is referred to as a "741-type." It is still the most popular type. The 747 is an early dual 741; the 748 is a 741 with the compensation capacitor omitted so it may be optimally compensated externally for gains greater than unity. The 771 series is an FET family that came much later than the number would indicate.

The 776/4250/3476 "programmable" type added a new dimension of versatility because all internal bias currents are proportional to a single external bias current. These devices may be operated much like a 741 at full current, as a lower-speed micropower op-amp at reduced current, or turned completely off. At the other extreme, the 791 power op-amp provides over one amp of output current. The remarkable thing is that other properties are

similar to a 741, except that compensation is external, including separate compensation for the output stage. There are many power op-amps available, but few others are monolithic.

The 101 was National's competitor to the 741/748. The 102/110 was faster, but it can be used only as a voltage follower. The 107 had reduced input current. The 108 employed "super-gain" or "superbeta" transistors for additional reduction. Back-to-back diodes between the inputs are included for protection; these may cause problems in some circuits. The 112 is a low-power version. The 116 used a Darlington input configuration for even further reduction at the expense of offset voltage.

The 118 was the first "high-speed 741." It is more likely to give problems than a 741, but this is usually true of a high-speed device. It is still the most common high-speed op-amp.

The 124, 148, 2902, 3503, and 4741 are all quad compensated op-amps; performance varies from slightly worse to slightly better than a 741. Actually the original quads were the 1900/3401 "Norton" or "current" op-amps. They are not true op-amps, as the inputs are controlled by currents and will not tolerate any significant voltage swing. There are few good reasons for using these. The LM192 was one of the first to offer different functions in the same package.

The 155, 156, and 157 are modern JFET-input op-amps. These introduced a new trend; instead of a single device, an entire family is introduced simultaneously, each device optimized for a specific purpose. The next four followed much later. The 13741 is an actual complete 741 with JFET followers added at the input to reduce current at the expense of offset voltage; it is intended as a "drop-in" replacement for existing 741 circuits. The LM10 was billed as a new-generation device. It had the low voltage offset of bipolar but input current rivaling that of an FET, worked at very low voltage, and threw in an accurate voltage reference for free. The LM12 is the latest offering, giving a whopping 10 amp output, unbelievable (at this time) for a monolithic.

The 1539 is an early design having high (at the time) bandwidth and slew rate. The 1558, 158, and 2904 are dual 741's. The MC3503 is billed as a "single-supply" quad. Of course, any op-amp may be configured to work from a single supply, but these types are designed to accommodate lower supply voltages and the working input range usually includes the negative supply voltage (normally ground).

The 3571 is a quad JFET-input. The 4202 and 146 are quad programmable (biasable) op-amps. Unfortunately, the four op-amps cannot be set individually because there are not enough pins in the standard package, and the partitioning and pinout are not yet standard for these.

The 909/911 was the original low-noise op-amp. The 2400 is a unique "programmable" quad; the term here means not biasable but selectable by a two-bit digital code. One op-amp is on and the other three off; the inputs are separate but the output is common.

The 2530 has both high slew rate and high bandwidth at the expense of not being a true op-amp; it must be used in the inverting configuration. The 2620 is the original "decompensated" op-amp, which implies the following: a compensated monolithic op-amp must be designed to be reasonably close to being stable without the compensation, because it is impractical to require a large capacitance in monolithic construction. Therefore, if the compensation is omitted, the device typically will be stable at some reasonable gain setting, for example 10. If no gain is tolerable, a resistor across the inputs can reduce the loop gain and fool the op-amp. This is generally acceptable, as the op-amp usually has excess gain for the unity-gain configuration. The 2900 is "chopper-stabilized." This type achieves very good input characteristics by taking the input dc and "chopping" or switching it (switches typically have much less offset than amplifiers) to create ac, amplifying the ac, and converting back to dc with another switch. This naturally introduces some noise, and some types are restricted to certain configurations. The 4602 is a "high-speed quad 741," being about an order of magnitude faster than a 741 even though it is completely compensated. The 4622 was a decompensated version; both are being discontinued. (The quads were really the driving force behind decompensation because a quad has no extra pins for external compensation.)

The 3060 and 3080 were the first "transconductance" amplifiers. These have a current output rather than voltage. This produces some minor advantages but also some minor disadvantages. The output typically swings close to the rails, but current is relatively low. Slew rate is good, especially at low bias currents, but does depend on loading. Theoretically, these devices can be used to do some nonlinear tasks, such as taking logarithms, but it is usually better to use an IC designed for the task. Best usage is just like op-amps. It may be surprising that a voltage-mode device can be replaced with a current device, but circuit performance depends, basically, on the feedback and not the device itself. The 3130/3140/3160 family was unique in having MOSFET input transistors which have virtually infinite input impedance. The only current is the leakage current of the zener diodes protecting the inputs; the input capacitance is about 4 pf. For comparison, this is equivalent to a 1-inch piece of 50-ohm coax. The 3130 and 3160 also have a CMOS output. This has the advantage of swinging completely from "rail-to-rail." It has the disadvantage that CMOS is limited to about 15V total supply voltage, but this becomes irrelevant if it is used single supply. It has an additional advantage that the output stage is linear; there is no crossover" distortion, which can cause problems in certain circuits requiring low distortion at high frequency.

The 4136 was the original quad true op-amp; unfortunately, it has an unusual pinout and, therefore, is not recommended. The TL061 series is a very complete family of JFET-input op-amps. This may be an indication of what the future holds; the number of devices will be so great that it is tedious even to list them.

The 761X, etc., series (x denotes another number, which defines some option) is a new class called CMOS op-amps. All transistors are MOS, usually occurring in complementary pairs. These work at very low supply voltages, 1V total here, which allows the use of single-cell batteries. They work well at low power. Input range may include both rails. Output range is normally rail-to-rail, which is particularly important at low supply voltage. The

usual requirement of swinging within a few volts of the rails is not good enough at supply voltages less than 5V.

The OPA111 is the lowest known noise of FET input types and also has excellent input specifications. The OP27/37 beats it in voltage noise, but current noise becomes a factor here, so which op-amp is best depends on impedances. The LT1028 is a recent entry having a better voltage noise but worse current noise. The NE5539 is probably the fastest available monolithic, with a gain-bandwidth product of 1.2 GHz and a slew rate of 600V/microsecond. These require extreme care in circuit design and layout, and restrictions apply. They should not be considered ordinary op-amps.

TROUBLESHOOTING

It is impossible to formulate a standard procedure for "debugging" op-amp circuits, as the variety of circuits has been shown to be very wide and likewise the number of things that can go wrong is unbelievably infinite. Nevertheless, there are a few common ailments and there is a sensible approach to follow when and if your op-amp circuit doesn't work. When a circuit fails to operate properly, the first impulse is to replace the op-amp. This is generally a bad idea, as the trouble is usually elsewhere and the op-amp is often soldered in. It makes sense to check the easy things first; Table 3 gives such a "checklist." The table is divided into nonlinear and linear categories. This separation is rather artificial and any procedure listed under one may apply at times to the other.

Nonlinear circuits are generally easier to troubleshoot, as there is less to go wrong. The first two items constitute a simple, visual inspection that may save a lot of time. All technicians can make mistakes, and some are even colorblind! Next, the supply voltages should be checked <u>on the pins of the op-amp</u> (comparator, really). They may be OK at the power supply or another point on the board, but they must reach the op-amp to do any good. This includes ground; make sure the ground in the vicinity of the op-amp is at zero volts and not something else. Next, check the signal voltages at the two op-amp inputs. If these two voltages are what they should be, then the output should be correct. Up to this point you have not unsoldered anything. If all the input conditions are proper but the output is not, then disconnect the load, as the output may be unable to assume the desired state. For a nonlinear circuit, the output should be either ±V; if it is neither, step 4 may be skipped. If disconnecting the load has no effect, replace the op-amp. While the op-amp is removed, check the voltages of <u>all</u> the pins, including the ones you <u>think</u> are not being used. You may have more inputs than you know about. The new op-amp will almost always either work properly, indicating a bad op-amp originally, or do the same thing as the original, indicating a problem in the rest of the circuit. If it does something different, then you have two problems instead of one.

A linear circuit will frequently malfunction by operating in a nonlinear mode. If so, refer to the procedures for nonlinear circuits. (Similarly, a nonlinear circuit found to be operating in a linear mode may require the linear troubleshooting procedures, but this is a rare occurrence. If the

circuit is operating linearly or partially linearly, steps 1-3 above still apply.) Step 4 may be tried, but it is usually difficult because the difference between the signal voltages at the two op-amp inputs is normally very small; if not, this is in itself indication of a problem. On op-amps where the output may be grounded safely, shorting the output will remove the effects of feedback and allow a large voltage between the inputs. The compensation pins (and any others available) give test points for observing some of the internal happenings of the op-amp. If the manufacturer gives an internal circuit diagram, these signals may give additional clues as to what is going on. If this fails, try steps 5 and 6 as for nonlinear. Another trick is to open feedback paths. Circuit operation will then be entirely different (i.e., nonlinear), but it may give some idea as to what is going on. If the op-amp appears simply to not be doing its job, recheck the specification sheet to make sure you are not exceeding its capabilities.

One class of problem happens so frequently in linear circuits that it deserves special mention. In addition to the desired signal, an unwanted signal or "noise" appears at the output. The first type of noise is hum, ripple, or high frequency "hash" picked up from another part of the circuitry. A frequent cause is ground leads shared with other circuitry which must be beefed up or separated. The first step should always be to put the scope probe on circuit ground. If the noise appears there, it will appear everywhere. Note that it may simply be a measurement problem, i.e., the scope is ungrounded or is grounded at a bad place. This author insists on use of a linear, high-speed scope. It tells the whole story. A meter gives only one number, which is usually an average. A digital scope rejects anything above its bandwidth and may or may not record transients. Another cause is common power supply leads, which may require the same precautions or else decoupling. A less likely cause is capacitive or inductive pickup due to proximity to other circuitry which may require shielding or more physical separation.

A second type of noise is a high frequency oscillation riding on or even obscuring the signal. The first and easiest step is to bypass power supply leads to ground with large capacitances to ensure that the dc supplies really are dc. Do it as near to the offending op-amp as possible. The next step is to remove the load, as it does change the op-amp characteristics somewhat. Frequently with reactive loads, a small (say 100 ohm) resistor is needed in between the op-amp and the load so the high frequency phase shift characteristics of the op-amp are not altered. It may also be necessary to increase the frequency compensation of the op-amp to account for stray capacitances. In frequency selective circuits, compensation should not play a part in the overall response, as otherwise changing the op-amp may change the frequency response. Generally, the circuit should be designed so that changing the compensation capacitors by a factor of 2 in either direction does not change the frequency response or cause oscillations. Circuit layout is seldom the cause, as op-amps are usually used at relatively low frequencies (1 MHz). An exception is high gain amplifier stages (including clippers) where the output leads must be kept away from the op-amp inputs. Several tricks used with PC (printed circuit) cards are: ground planes in every other rack slot, large ground strips between leads on the PC board, and double-sided boards where one entire side is a ground plane.

Less likely is an oscillation at relatively low frequency. First, recheck the circuit design to make sure there is not an unstable loop, not only in normal linear operation but in case an op-amp accidentally gets driven out of its linear range. Otherwise, the most common effect is "motorboating" in high gain amplifiers which may require stiffer power supplies, beefier power supply leads, or decoupling in multi-stage circuits.

WHAT NEXT?

With the advent of the 741 and similar types, a general purpose, "black-box" device had pretty well arrived. The device was insensitive to supply voltage, inputs and outputs were protected, and compensation was internal. The device could be used "as is" in the majority of applications. The major effort thereafter was in the improvement of specific parameters. Orders-of-magnitude improvements have been made in many areas. This was first true for special-purpose devices, where some parameter was improved at the expense of others. However, the situation has reversed, and new devices are usually complete improvements. For example, the new FET-input op-amps, in spite of having miniscule input current, are good general-purpose devices, faster than a 741, with offset voltage no worse, and not significantly more expensive.

A second thrust was to add new features to the basic op-amp concept to make it more versatile. The principal achievement here was the "programmable" op-amps. Little more has been done but, indeed, the op-amp is already extremely versatile, making it difficult to improve.

A major achievement was the advent of quads. This may not appear significant, but it reduces the power supply wiring 75% and the area 50% and at virtually no loss in performance. Furthermore, the quads are not much higher in price and supply current than the old singles!

Another achievement was the true "single-supply" op-amp. These will do everything the original op-amps did but, in addition, will perform fully with the inputs at negative supply voltage (i.e., ground) and operate at much lower supply voltage. In fact, there are several designs which will operate from a single battery cell.

Some prewired circuits using op-amps have appeared, notably instrumentation amps, state-variable filters, and generalized impedance converters. Impetus in this area has been partially nullified by improvements in ease of use of basic op-amps themselves; for example, the difference between a state-variable filter and a quad op-amp is only a few passive components, which the user might wish to select anyway. A relatively new trend is the inclusion of extra circuitry (e.g., voltage reference) or mixing of functions (op-amp/comparator). Manufacturers are getting better at mixing analog and digital, so many of the new functions are digital (e.g., digital gain set).

Now we come to the part where most writers get into trouble--predictions for the future. The trends described above will surely continue, with obvious exceptions, where a limit has already been reached; for example, the number of op-amps in a package has already reached the pin limit, except

possibly for prewired followers. More circuit blocks will become available, many utilizing op-amps internally. The op-amp will continue to be the work-horse, but the trend toward integrating entire subsystems will continue. Speed will increase; this is probably the most serious limitation of present-day op-amps, and the pressure for faster, higher-frequency systems is considerable. Power consumption will continue to dwindle. In general, devices will continue to creep toward that unattainable goal, the ideal op-amp.

CONCLUSION

Each edition of this primer gets longer and more involved. IC technology is expanding at a rapid rate, however, and the sophistication of systems is increasing commensurably. It is no longer adequate to be able to design a summer and an integrator. Circuits that were curiosities 10 years ago are in common use today.

The point of this report has been to acquaint the reader with the almost limitless uses of op-amps and general op-amp circuit design procedures, not to be a cookbook for ready-to-go circuits. A standard circuit cannot possibly cover all applications and may get the user into much trouble if he attempts to use it without understanding the inherent design considerations. However, the manufacturers' application notes do frequently give complete circuits, and these may sometimes be used with minor modifications or none at all.

It has been an inherent purpose of this paper to indicate that any circuit or equation, no matter how formidable in appearance, is simple if one really understands it. The vast majority of op-amp circuits may be understood, in principle if not in actual numbers, upon inspection. Actual calculations rarely take more than a page (excluding errors). By now the reader should have grasped the paradox that although the exact, complete mathematical description of a real op-amp circuit may be quite involved, in the vast majority of engineering applications a much simpler equation suffices, usually the simple ideal equations given here.

It is hoped that two central ideas have been impressed. The first is that the op-amp by its nature is an extremely versatile device. The second is that through silicon semiconductor integrated circuit technology the op-amp has been developed to an extremely practical component.

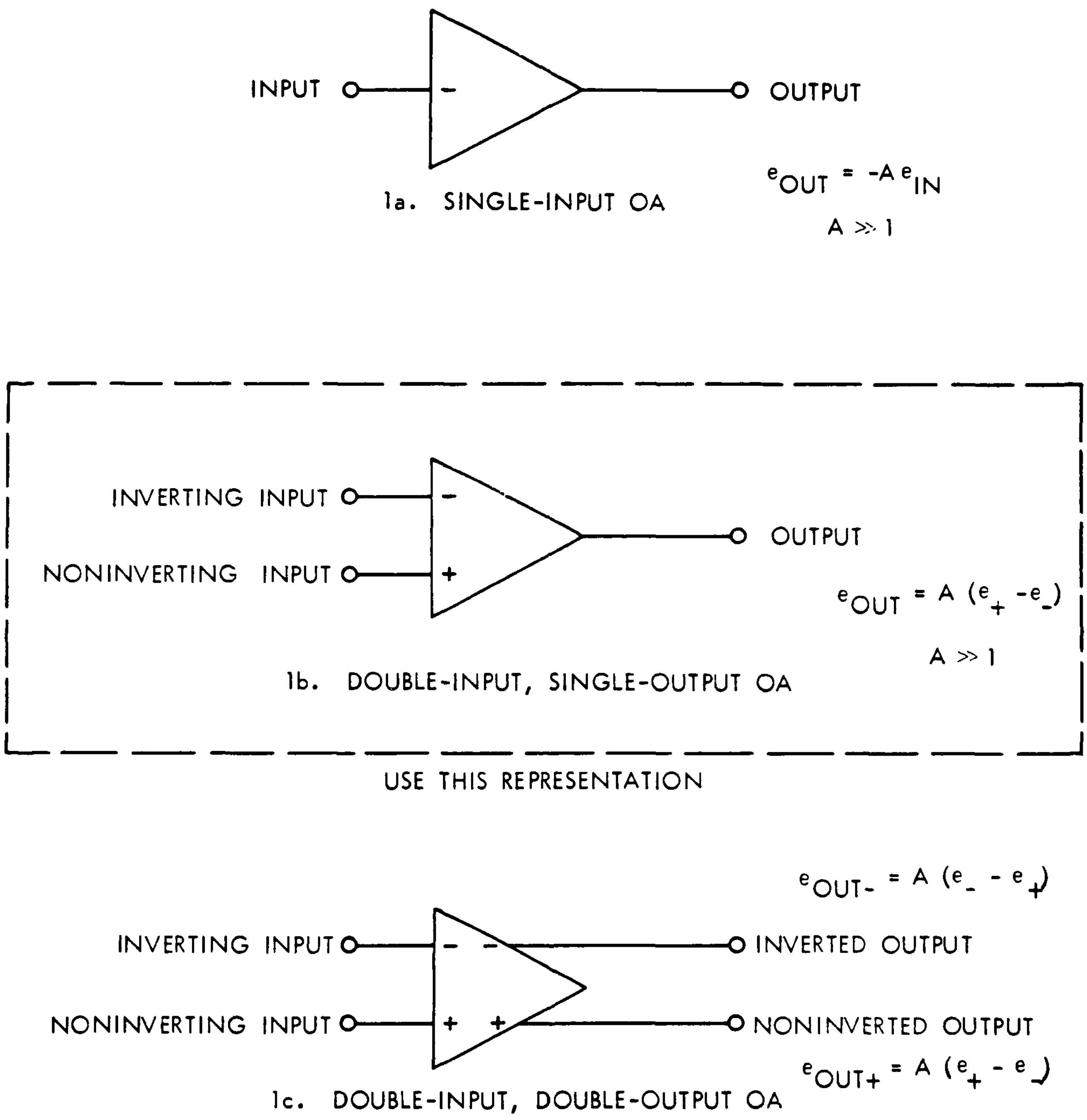
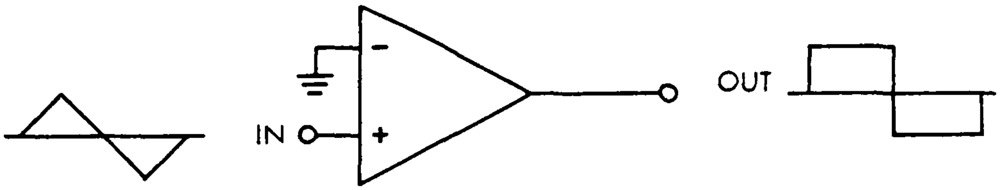
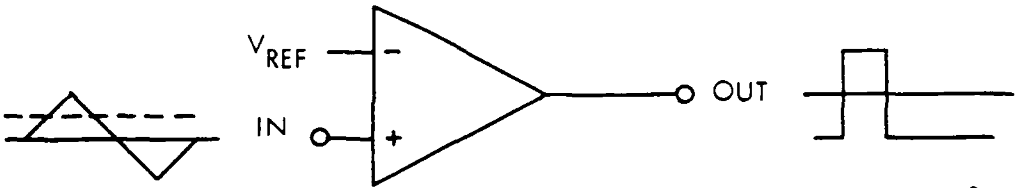


FIGURE 1. OPERATIONAL AMPLIFIER SYMBOLS



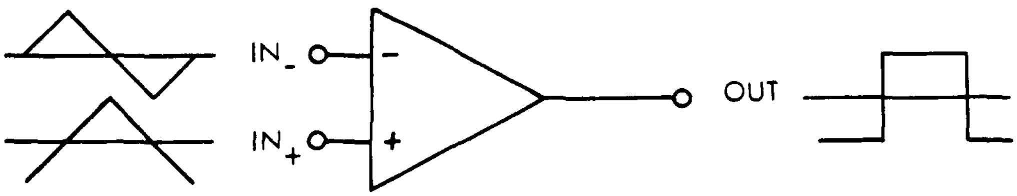
2a. CLIPPER

$$e_{OUT} = \begin{cases} +V; e_{IN} = + \\ -V; e_{IN} = - \end{cases}$$



2b. THRESHOLD DETECTOR

$$e_{OUT} = \begin{cases} +V; e_{IN} > V_{REF} \\ -V; e_{IN} < V_{REF} \end{cases}$$



2c. COMPARATOR

$$e_{OUT} = \begin{cases} +V; e_{+} > e_{-} \\ -V; e_{+} < e_{-} \end{cases}$$

FIGURE 2. BASIC NONLINEAR CIRCUITS

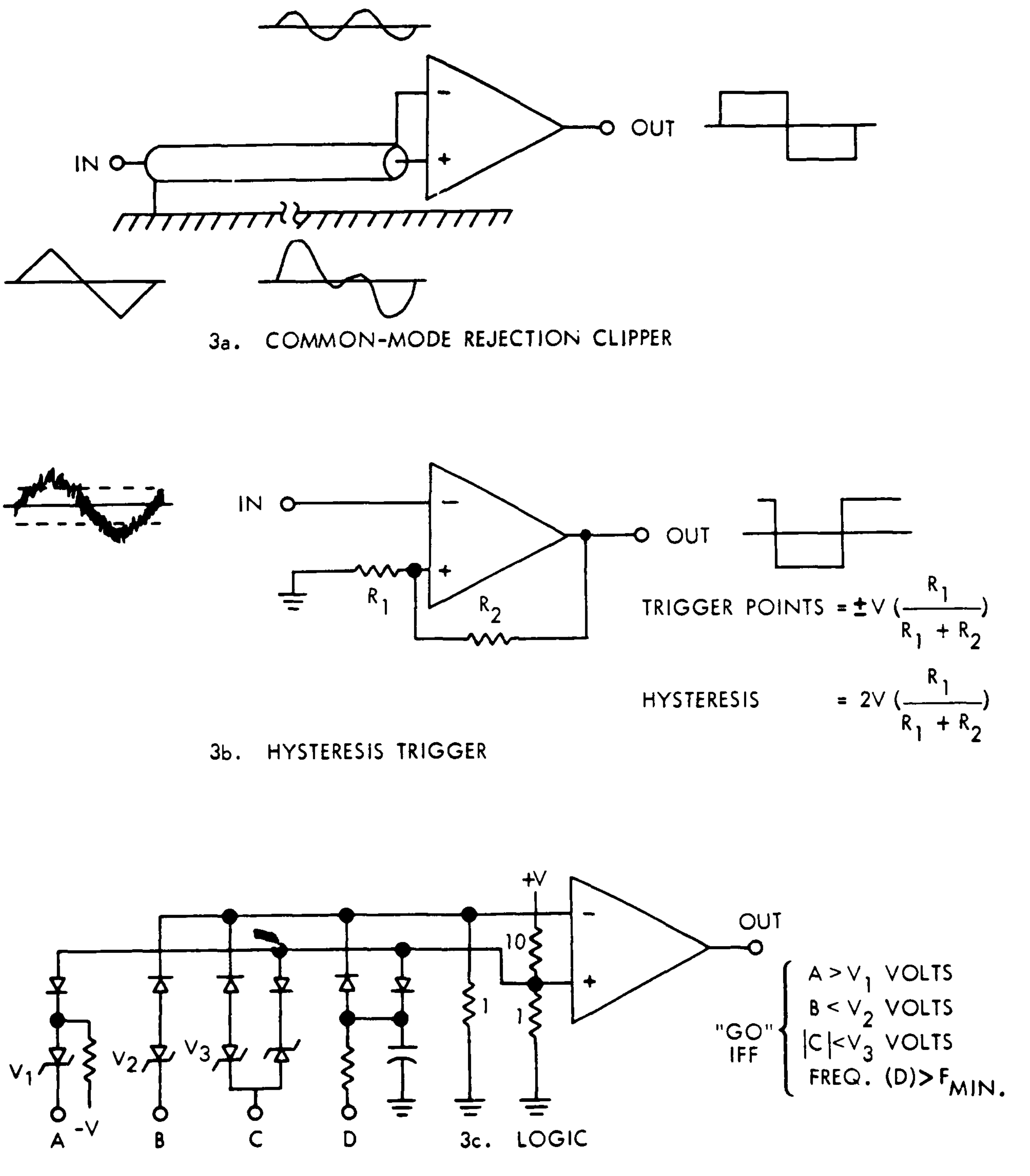
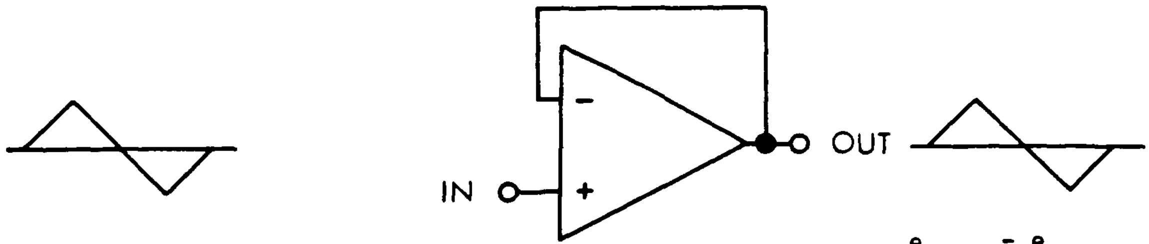
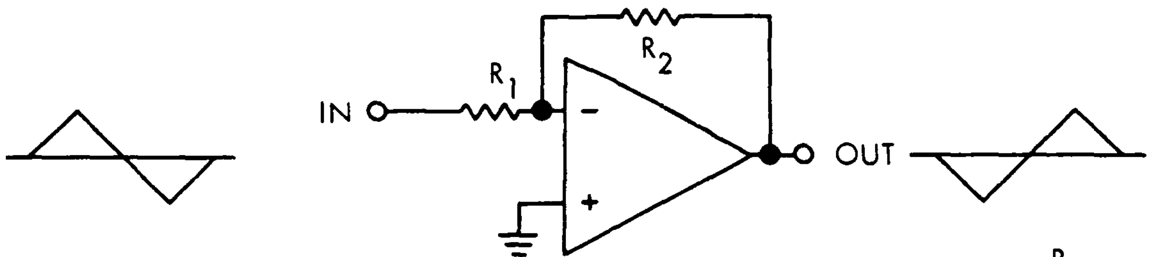


FIGURE 3. SPECIAL NONLINEAR CIRCUITS



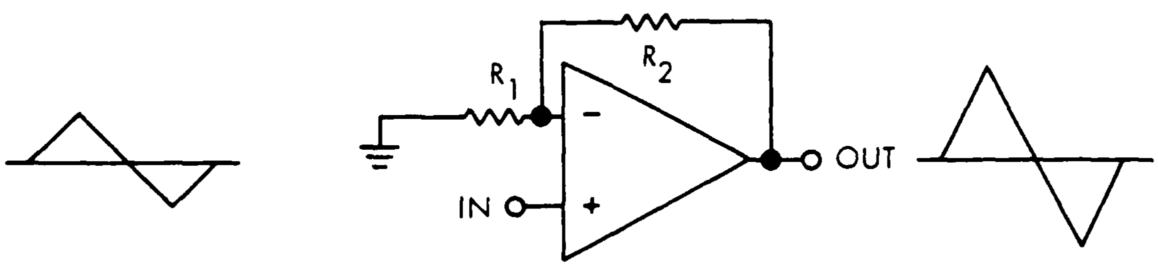
$e_{OUT} = e_{IN}$

4a. FOLLOWER



$e_{OUT} = -\frac{R_2}{R_1} e_{IN}$

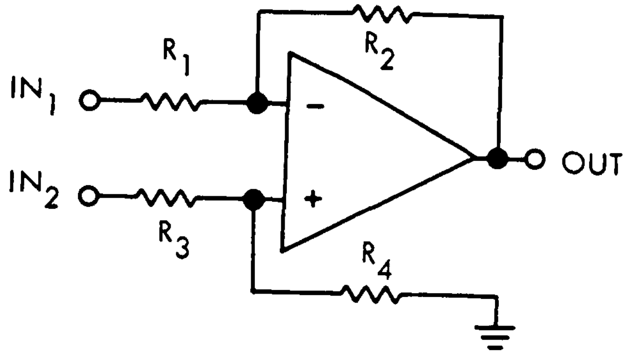
4b. INVERTER



$e_{OUT} = (1 + \frac{R_2}{R_1})\ e_{IN}$

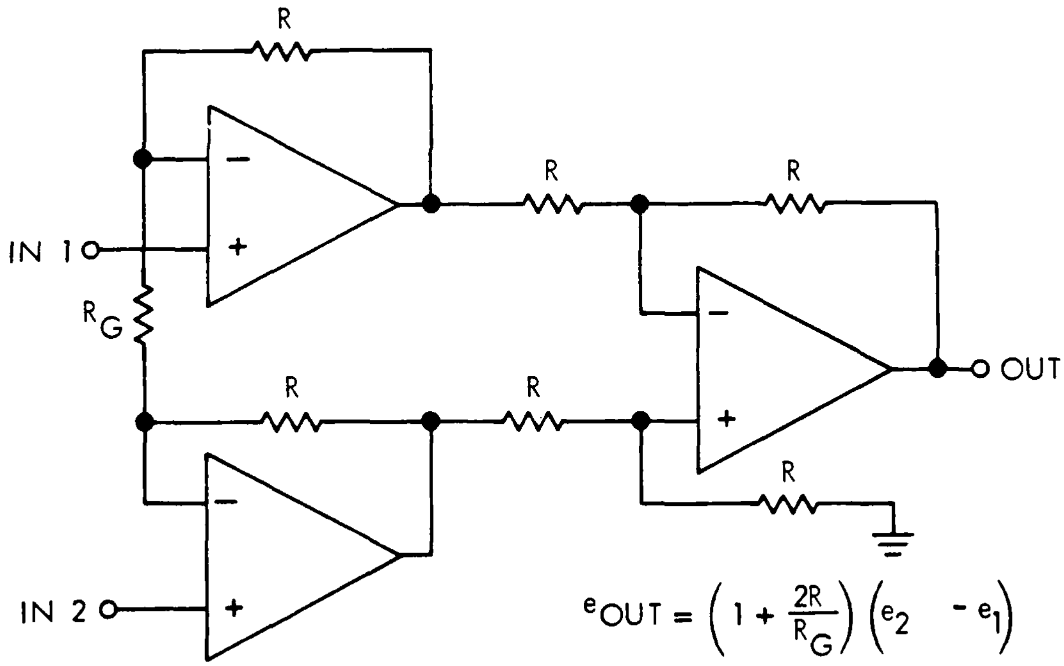
4c. NON-INVERTING AMPLIFIER

FIGURE 4. BASIC LINEAR CIRCUITS



$$e_{OUT} = \left(1 + \frac{R_2}{R_1}\right)\left(\frac{R_4}{R_3 + R_4}\right) e_2 - \left(\frac{R_2}{R_1}\right) e_1$$

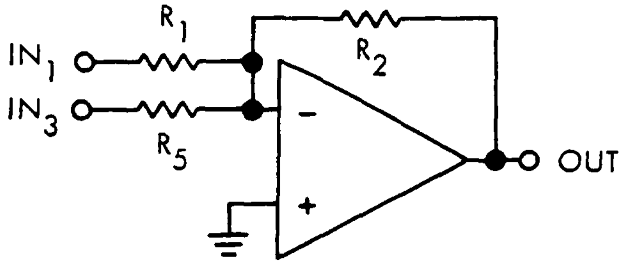
5a. DIFFERENCE AMPLIFIER



$$e_{OUT} = \left(1 + \frac{2R}{R_G}\right)\left(e_2 - e_1\right)$$

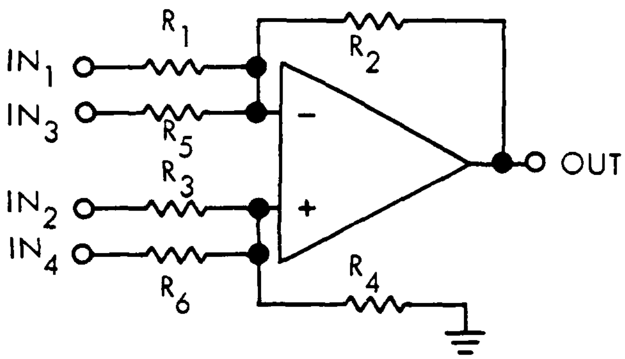
5b. INSTRUMENTATION AMPLIFIER

FIGURE 5. MULTIPLE-INPUT AMPLIFIERS



$$e_{OUT} = -\left(\frac{R_2}{R_1}\right) e_1 - \left(\frac{R_2}{R_5}\right) e_3$$

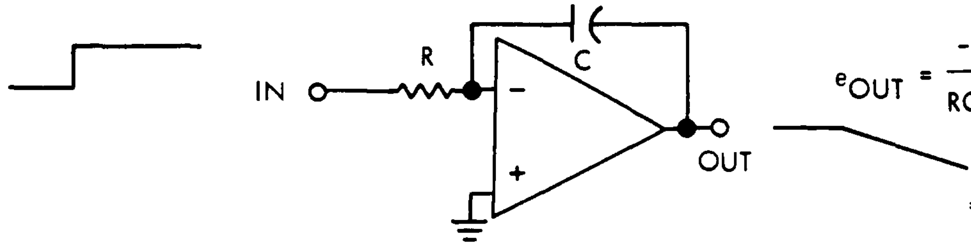
5c. SUMMING INVERTER



$$e_{OUT} = \left\{ 1 + \frac{R_2}{\dfrac{R_1 R_5}{R_1 + R_5}} \right\} \left\{ \frac{e_2}{1 + R_3 \dfrac{R_4 + R_6}{R_4 R_6}} + \frac{e_4}{1 + R_6 \dfrac{R_3 + R_4}{R_3 R_4}} \right\} - \left\{ \frac{R_2}{R_1} e_1 + \frac{R_2}{R_5} e_3 \right\}$$

5d. GENERAL AMPLIFIER

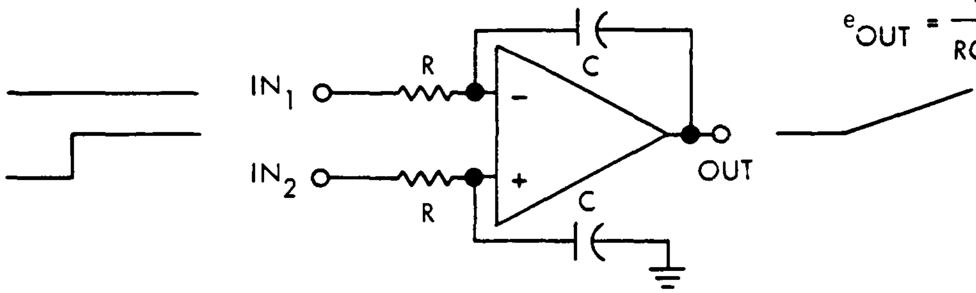
FIGURE 5. (Cont.)



$$e_{OUT} = \frac{-1}{RCS} e_{IN} \rightarrow \frac{-1}{RCj\omega} e_{IN} = \frac{-1}{\tau j\omega} e_{IN}$$

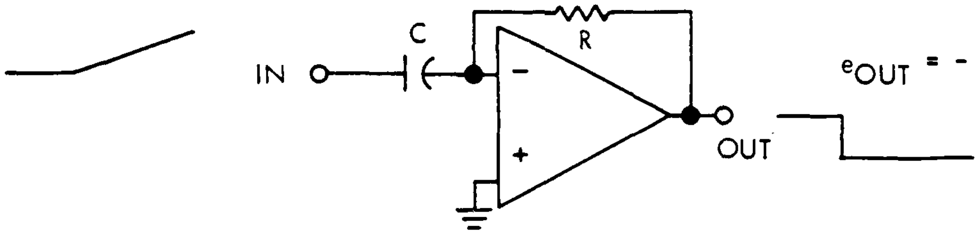
$$= \frac{-1}{RC} \int e_{IN} + C$$

6a. INVERTING INTEGRATOR



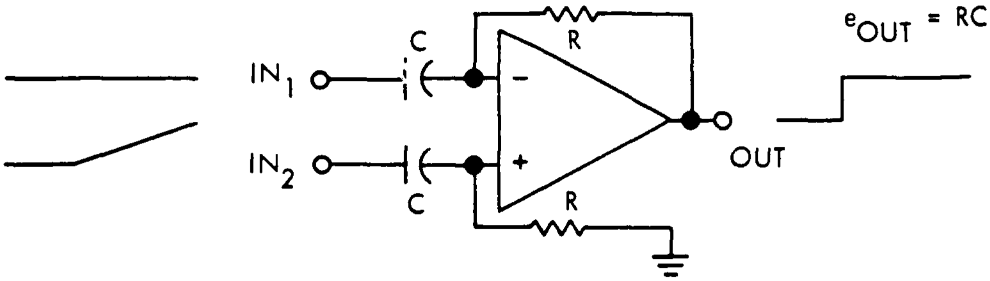
$$e_{OUT} = \frac{1}{RCS} e_2 - \frac{1}{RCS} e_1$$

6b. POSITIVE INTEGRATOR



$$e_{OUT} = - RCS\, e_{IN}$$

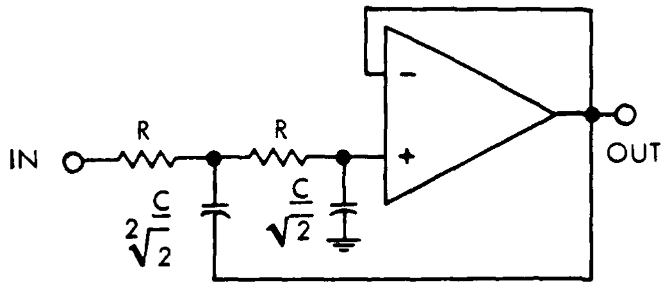
6c. INVERTING DIFFERENTIATOR



$$e_{OUT} = RCS\, e_2 - RCS\, e_1$$

6d. POSITIVE DIFFERENTIATOR

FIGURE 6. BASIC FREQUENCY-SENSITIVE CIRCUITS

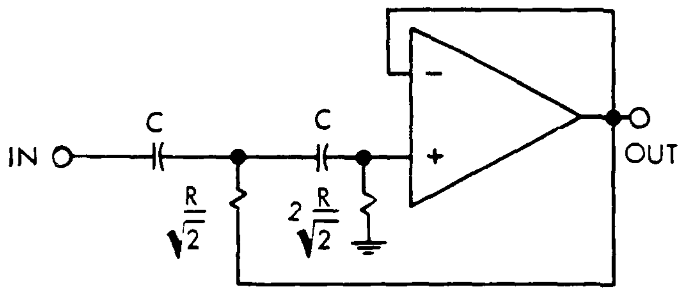


$$f_{CUTOFF} = \frac{1}{2\pi RC} = \frac{1}{2\pi\tau}$$

$$H(j\omega) = \frac{1}{\tau^2 (j\omega)^2 + \sqrt{2}(j\omega)\tau + 1}$$

$$|H(j\omega)| = \frac{1}{\sqrt{1 + \tau^4\omega^4}}$$

7a. UNITY-GAIN LOW-PASS FILTER (BUTTERWORTH)

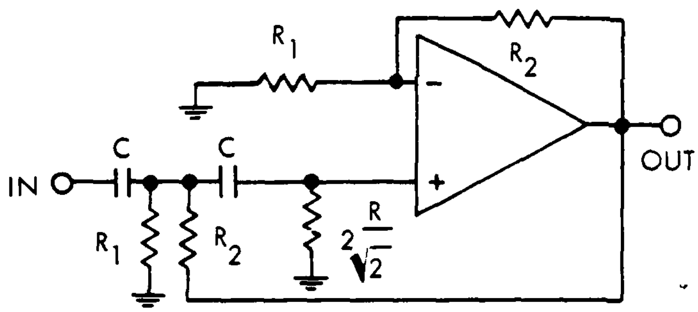


$$f_{CUTOFF} = \frac{1}{2\pi RC} = \frac{1}{2\pi\tau}$$

$$H(j\omega) = \frac{\tau^2(j\omega)^2}{\tau^2 (j\omega)^2 + \sqrt{2}(j\omega)\tau + 1}$$

$$|H(j\omega)| = \frac{\tau^2\omega^2}{\sqrt{1 + \tau^4\omega^4}}$$

7b. UNITY-GAIN HIGH-PASS FILTER (BUTTERWORTH)

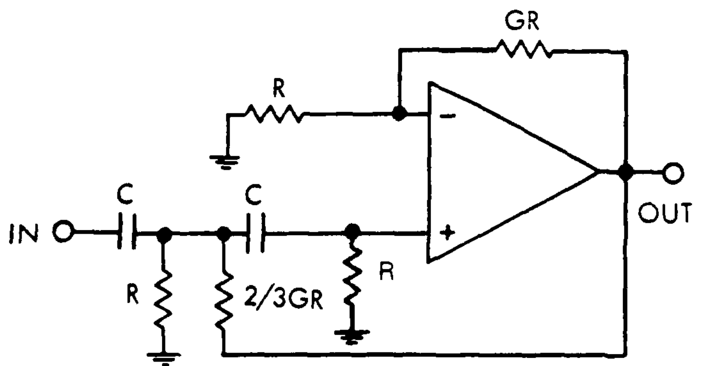


$$f_{CUTOFF} = \frac{1}{2\pi RC} = \frac{1}{2\pi\tau}$$

$$H(j\omega) = G\frac{\tau^2\omega^2}{\sqrt{1 + \tau^4\omega^4}}$$

7c. HIGH-PASS FILTER (BUTTERWORTH)

$$R_1 || R_2 = \frac{R}{\sqrt{2}}; G = 1 + \frac{R_2}{R_1}$$



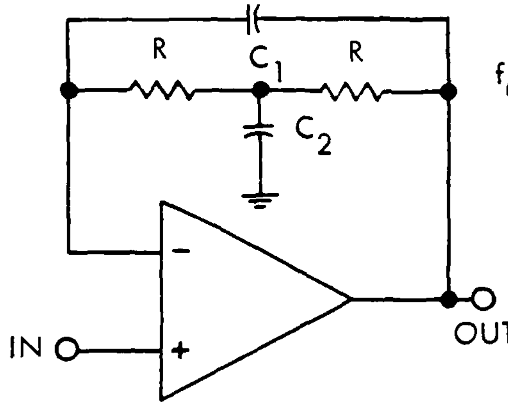
$$f_{CUTOFF} \quad \frac{1}{2\pi RC} = \frac{1}{2\pi\tau}$$

$$|H(j\omega)| \approx G\frac{\tau^2\omega^2}{\sqrt{1 + \tau^4\omega^4}}$$

$G \geq 15$

7d. HIGH-PASS FILTER (APPROX. BUTTERWORTH)

FIGURE 7. FILTER CIRCUITS

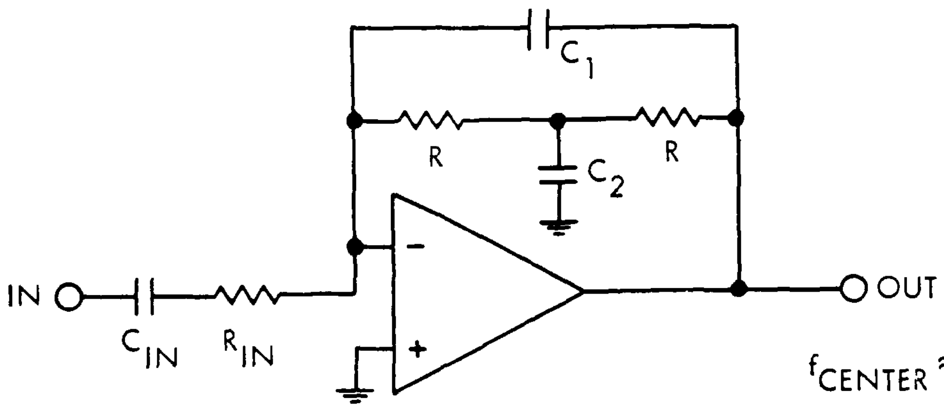


$$f_{CENTER} = \frac{1}{2\pi\sqrt{RC_1\,RC_2}}$$

$$Q = \frac{1}{2}\sqrt{\frac{C_2}{C_1}}$$

$$GAIN_{f_{CENTER}} = \frac{1}{2}\frac{C_2}{C_1} = 2Q^2$$

8a. NARROW BAND-PASS

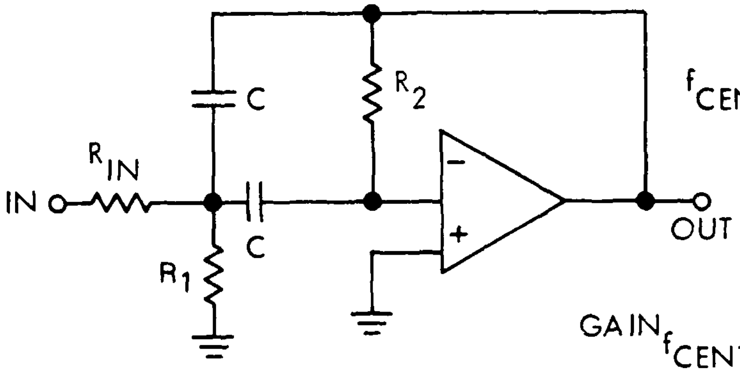


$$f_{CENTER} \approx \frac{1}{2\pi\sqrt{RC_1\,RC_2}}$$

$$Q \approx \frac{1}{2}\sqrt{\frac{C_2}{C_1}}$$

8b. IMPROVED NARROW BANDPASS

FIGURE 8. NARROW BAND-PASS FILTERS

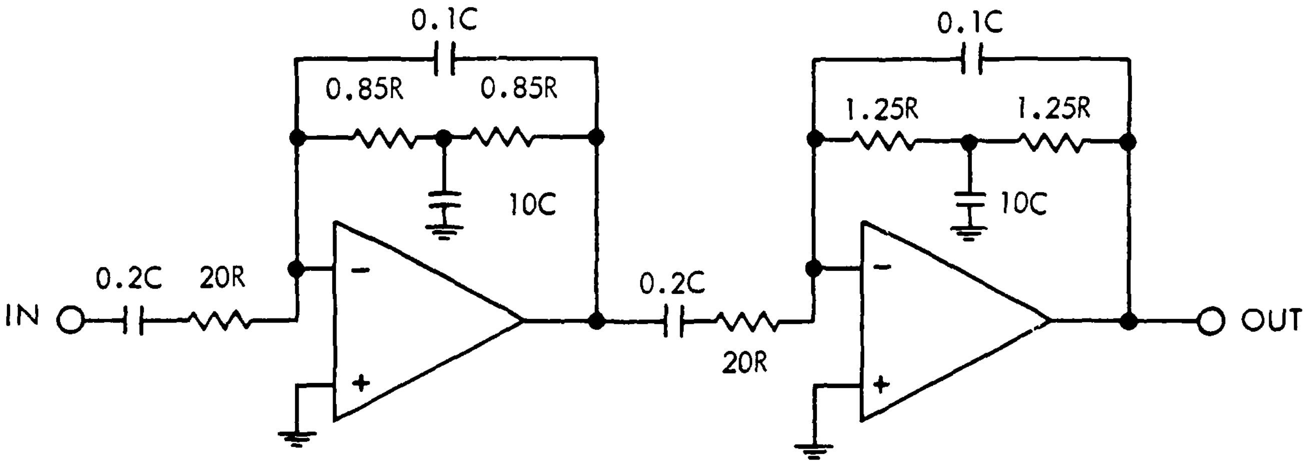


$$f_{CENTER} = \frac{1}{2\pi\sqrt{R_1C\,R_2C}}$$

$$Q = \frac{1}{2}\sqrt{\frac{R_2}{R_1}}$$

$$GAIN_{f_{CENTER}} \approx \frac{1}{2}\,\frac{R_2}{R_{IN}}$$

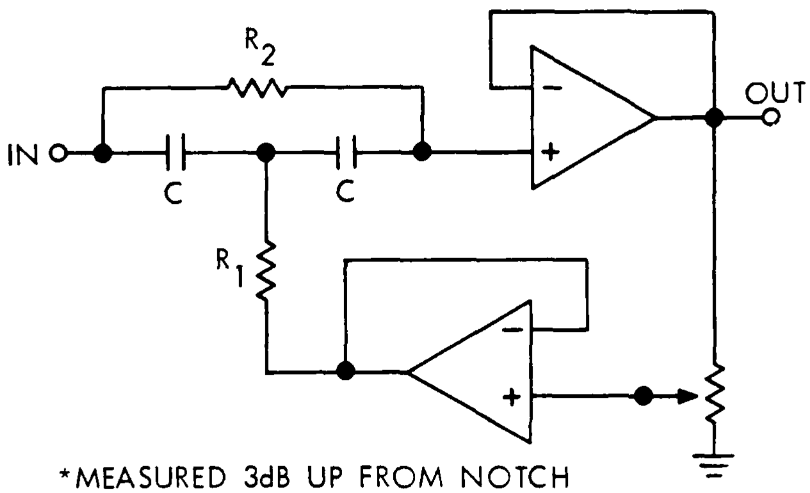
8c. ALTERNATE NARROW BAND-PASS



$$f_{CENTER} = \frac{1}{2\pi RC}$$

8d. OCTAVE BAND-PASS

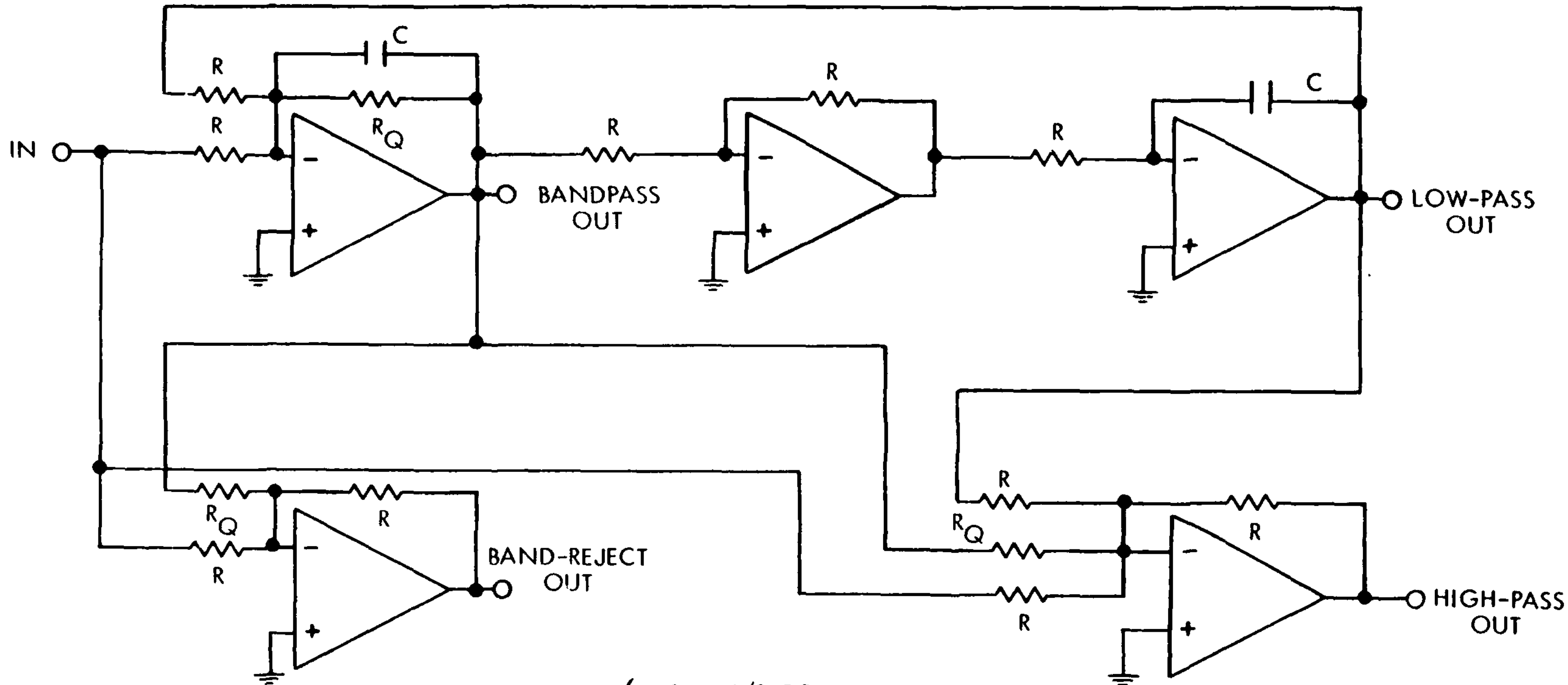
FIGURE 8. (Cont.)



$$f_{CENTER} = \frac{1}{2\pi\sqrt{R_1 C R_2 C}}$$

$$Q^* = \frac{1}{2}\sqrt{\frac{R_2}{R_1}}$$

FIGURE 9. NARROW BAND-REJECT FILTER



$$\left(\frac{e_{OUT}}{e_{IN}}\right)_{BP} = -\frac{S/RC}{S^2 + S/R_QC + 1/(RC)^2} \qquad \begin{cases} fo = 1/2\pi RC \\ G_{fo} = R_Q/R \\ Q = R_Q/R \end{cases}$$

$$\left(\frac{e_{OUT}}{e_{IN}}\right)_{BR} = -\frac{S^2 + 1/(RC)^2}{S^2 + S/R_QC + 1/(RC)^2} \qquad \begin{cases} fo = 1/2\pi RC \\ G_{fo} = 0 \\ Q = R_Q/R \end{cases}$$

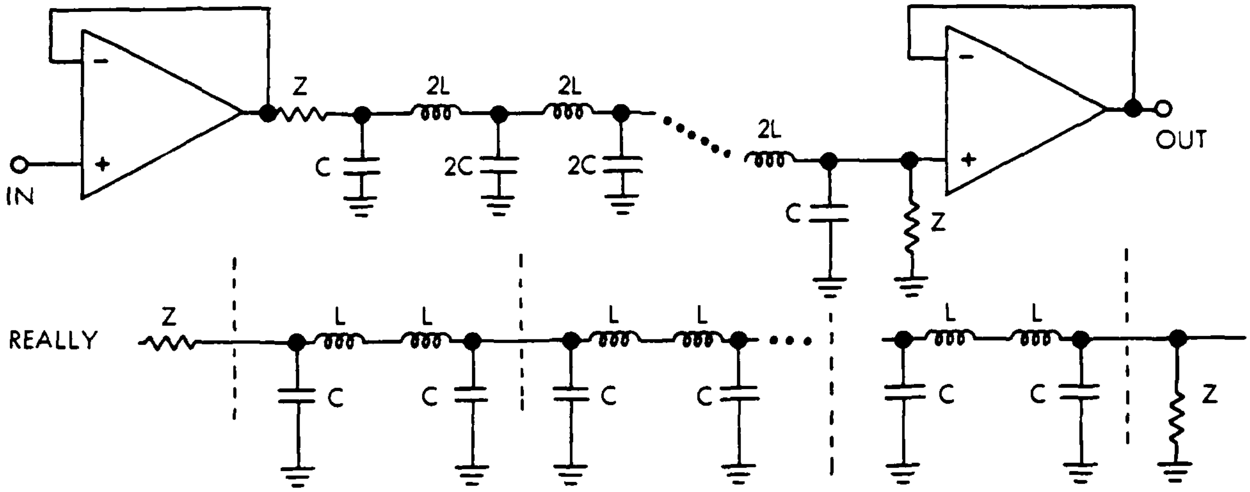
$$\left(\frac{e_{OUT}}{e_{IN}}\right)_{LP} = -\frac{1/(RC)^2}{S^2 + S/R_QC + 1/(RC)^2} \qquad \begin{cases} \text{FOR CRITICAL } \underline{\text{FREQUENCY}} \text{ DAMPING} \\ R_Q = R/\sqrt{2} \\ G_{fo} = R_Q/R = 0.707 \text{ (3db)} \end{cases}$$

$$\left(\frac{e_{OUT}}{e_{IN}}\right)_{HP} = -\frac{S^2}{S^2 + S/R_QC + 1/(RC)^2} \qquad \begin{cases} \text{FOR CRITICAL } \underline{\text{FREQUENCY}} \text{ DAMPING} \\ R_Q = R/\sqrt{2} \\ G_{fo} = R_Q/R = 0.707 \text{ (3db)} \end{cases}$$

NOTE:
FOR CRITICAL DAMPING IN TIME DOMAIN $R_Q = R/2$

FIGURE 10. STATE-VARIABLE FILTER

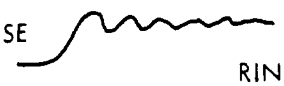
11a. LOW-PASS "DELAY LINE" FILTER



MATCHING IMPEDANCE $Z = \sqrt{\frac{L}{C}}$

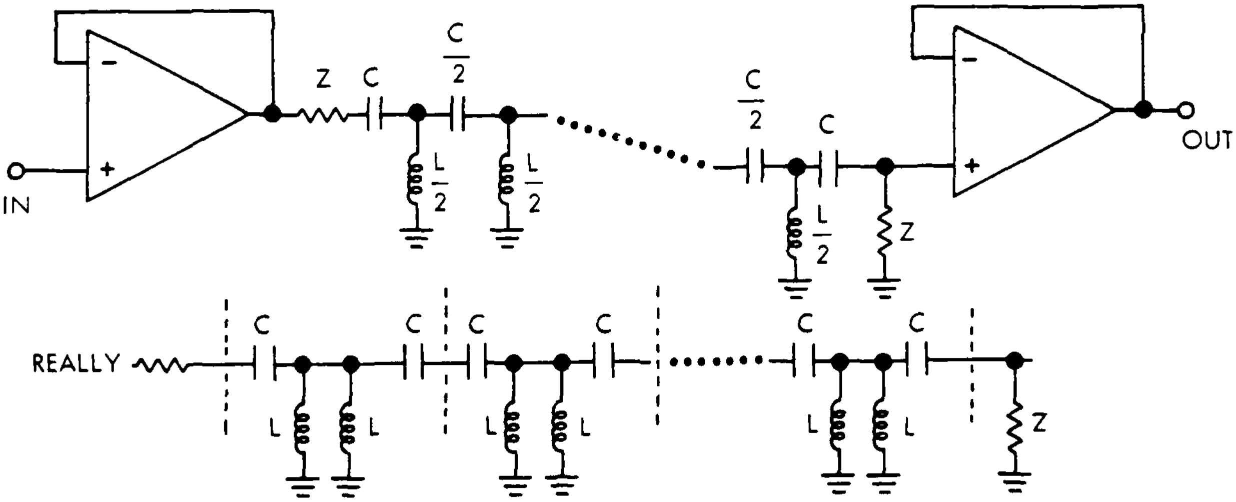
DELAY PER HALF-SECTION $d' = \sqrt{LC}$

CUTOFF FREQUENCY (3db) $f_c = \frac{1}{2\pi\sqrt{LC}}$

STEP RESPONSE 

RINGING FREQUENCY = f_c

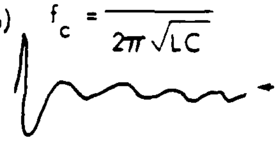
11b. HIGH-PASS "ADVANCE (?) LINE" FILTER



MATCHING IMPEDANCE $Z = \sqrt{\frac{L}{C}}$

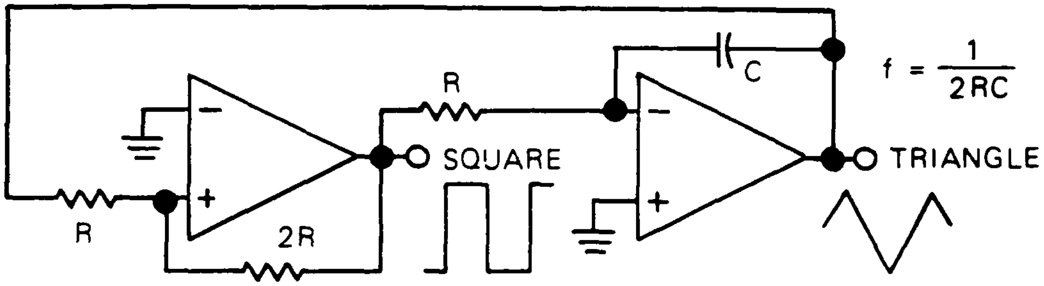
CUTOFF FREQUENCY (3db) $f_c = \frac{1}{2\pi\sqrt{LC}}$

STEP RESPONSE

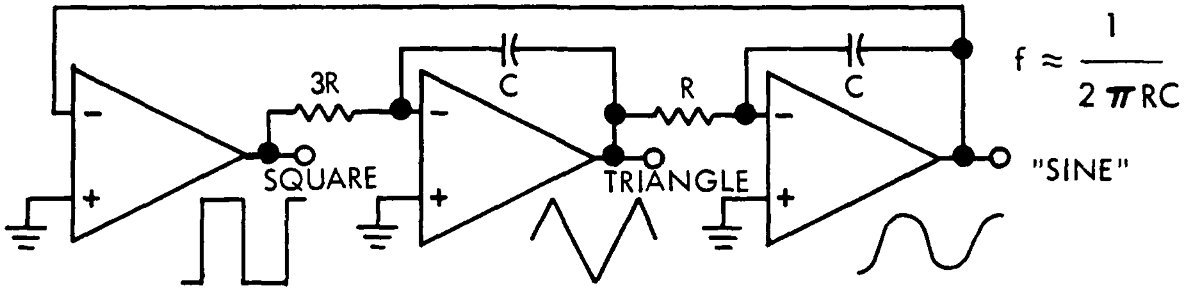


SETTLES TO RINGING FREQUENCY = f_c

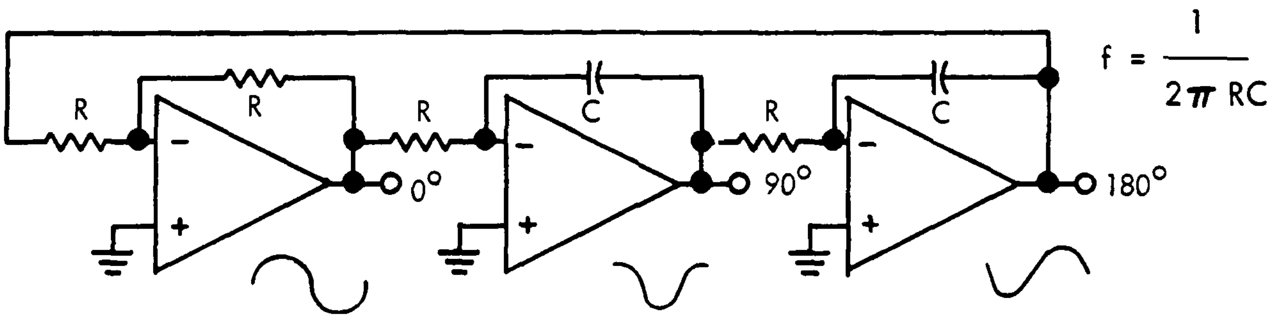
FIGURE 11. PASSIVE FILTERS



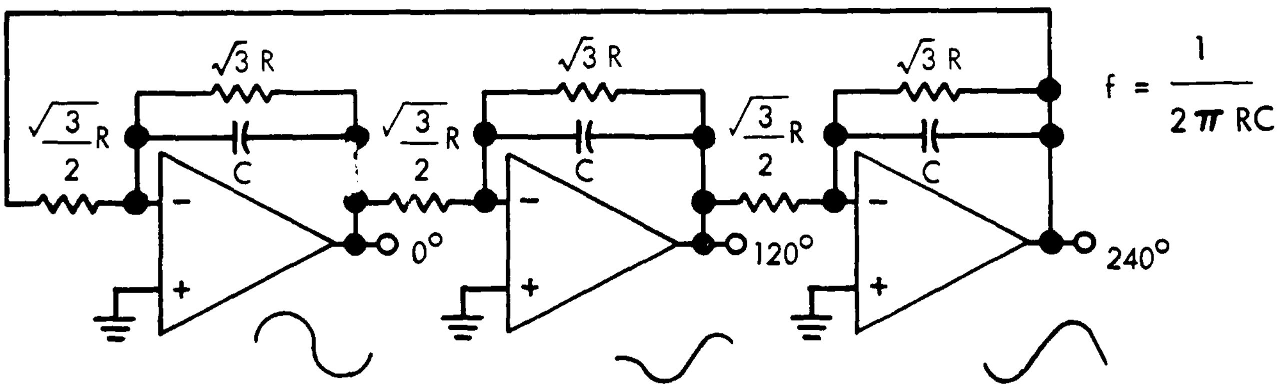
12a. TRIANGLE GENERATOR



12b. FUNCTION GENERATOR

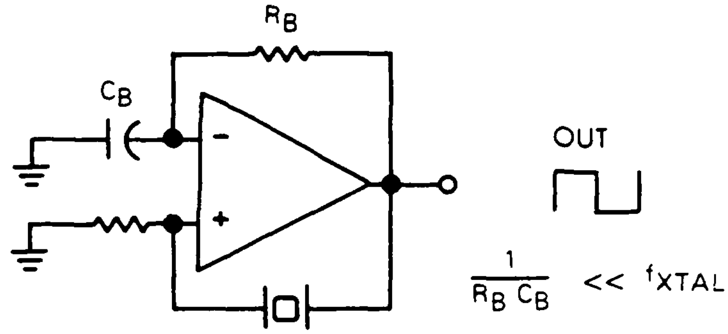


12c. QUADRATURE OSCILLATOR

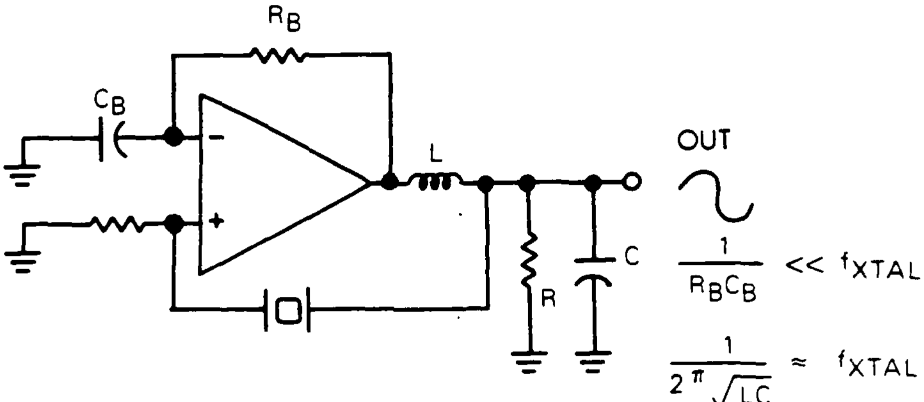


12d. THREE-PHASE OSCILLATOR

FIGURE 12. RC OSCILLATORS

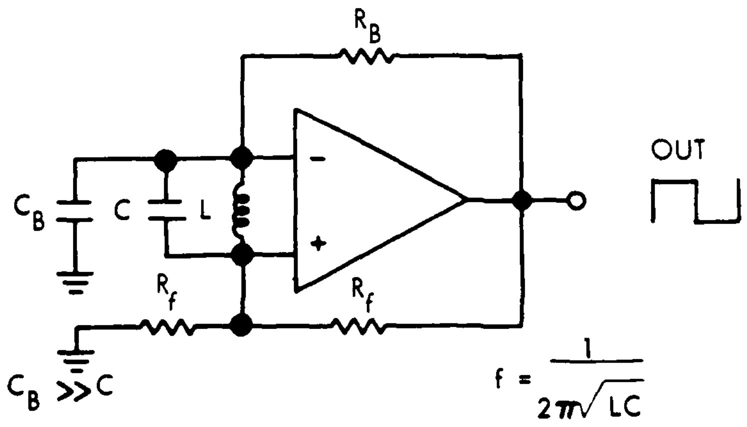


13a. SQUARE-WAVE CRYSTAL OSCILLATOR



13b. "SINE"-WAVE CRYSTAL OSCILLATOR

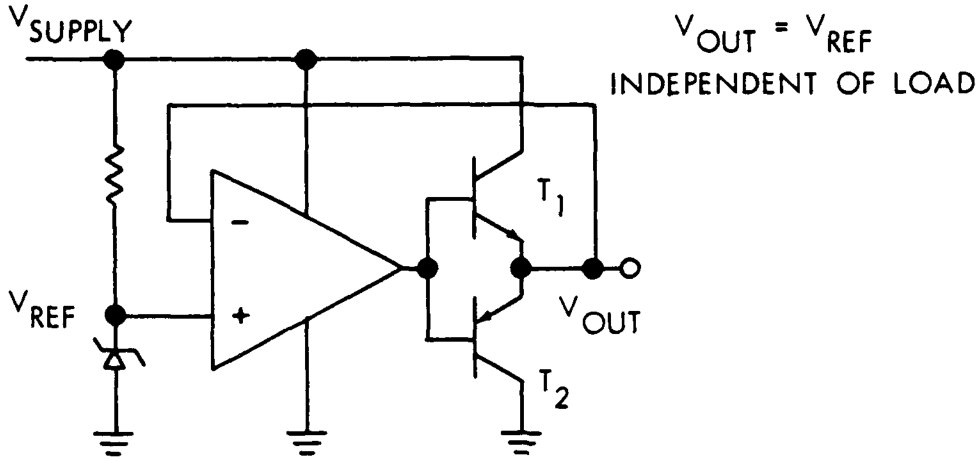
$R \approx \sqrt{\frac{L}{C}}$



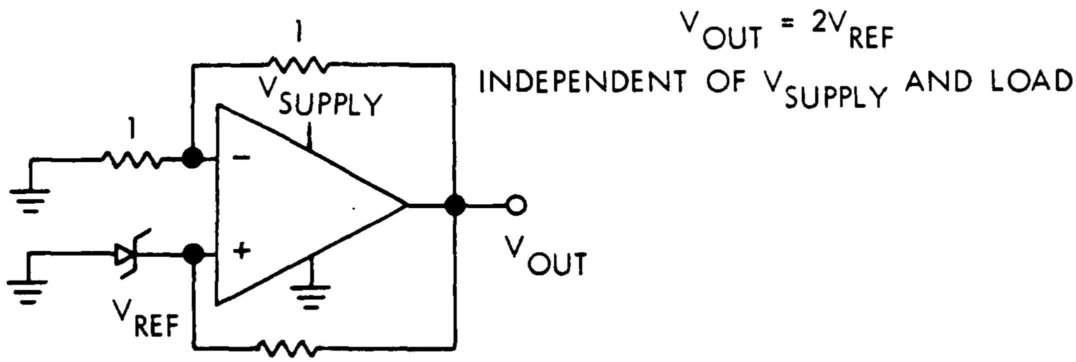
13c. LC OSCILLATOR

$$\frac{1}{R_B C_B} << \frac{1}{2\pi\sqrt{LC}}$$

FIGURE 13. CRYSTAL AND LC OSCILLATORS

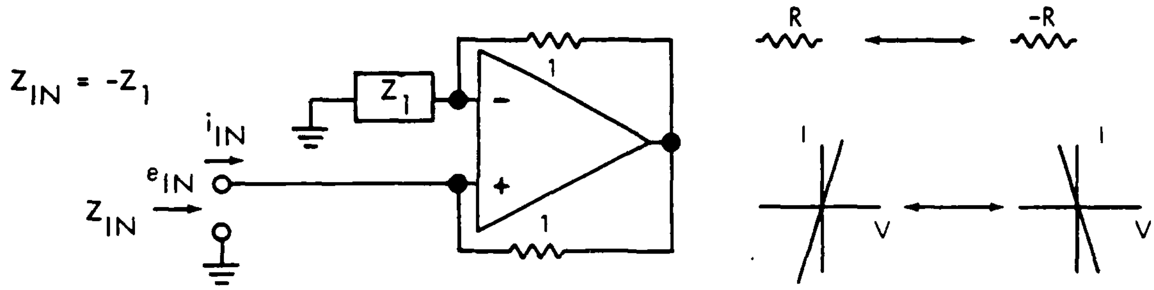


14a. BUFFERED REGULATOR

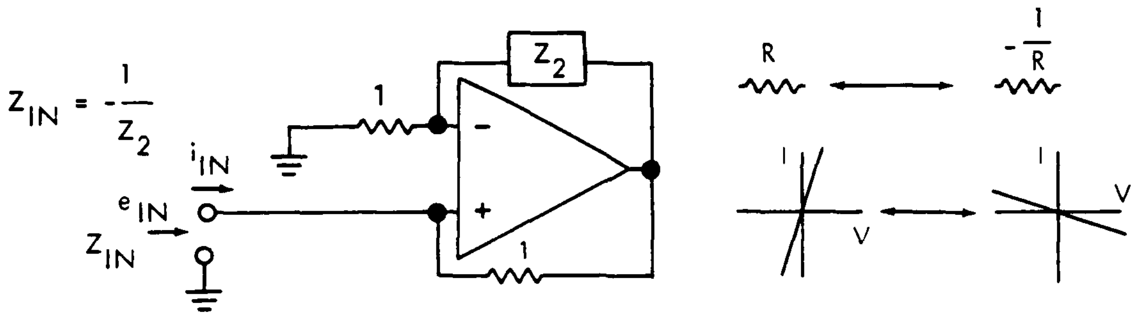


14b. BOOTSTRAPPED REGULATOR

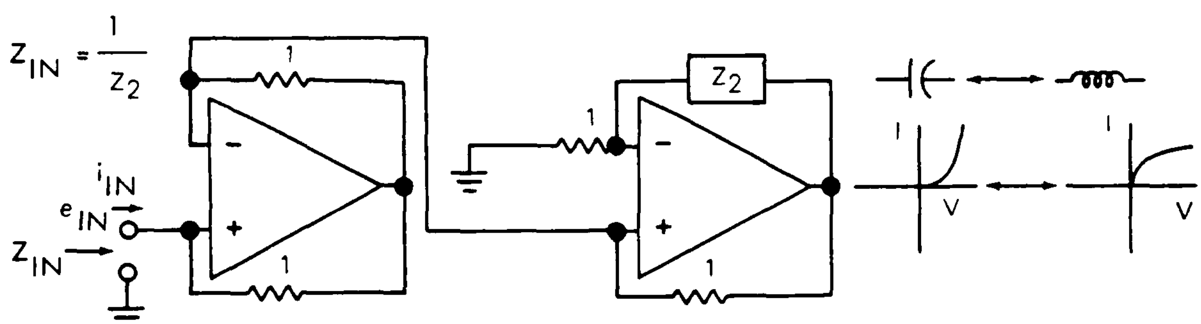
FIGURE 14. VOLTAGE REGULATORS



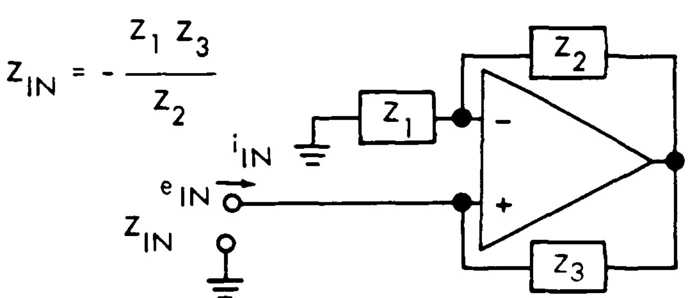
15a. NEGATIVE IMPEDANCE CONVERTER



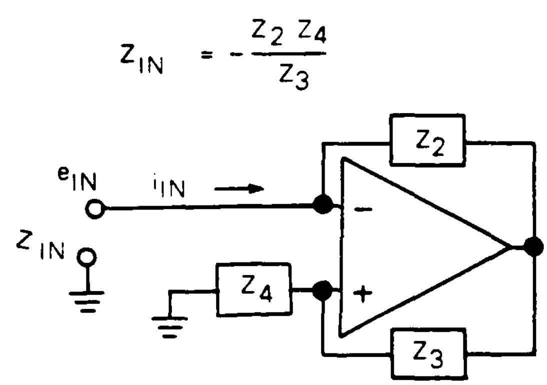
15b. NEGATIVE GYRATOR



15c. POSITIVE GYRATOR



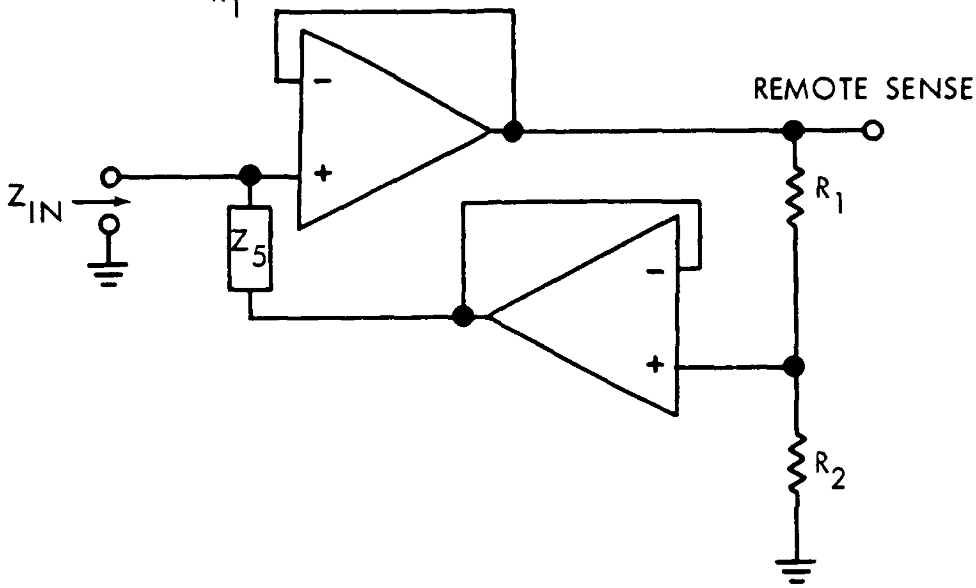
15d GENERAL CIRCUIT



15e ALTERNATE

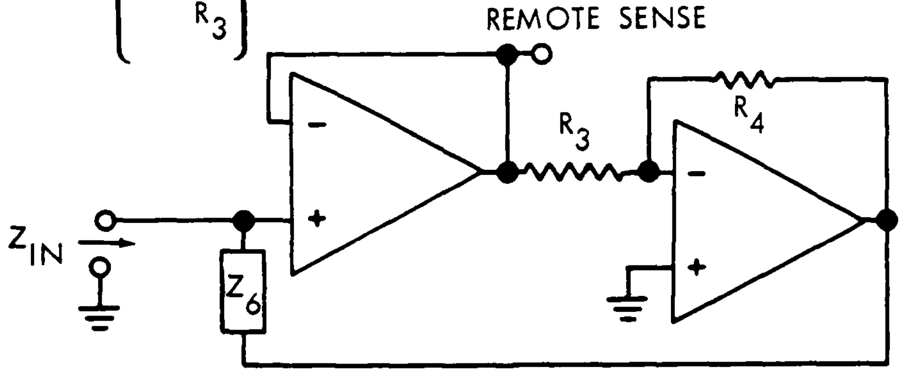
FIGURE 15. IMPEDANCE TRANSFORMING CIRCUITS

$$Z_{IN} = Z_5 \left(1 + \frac{R_2}{R_1}\right) = \frac{Z_5}{a} \; ; \; a = \frac{R_1}{R_1 + R_2}$$



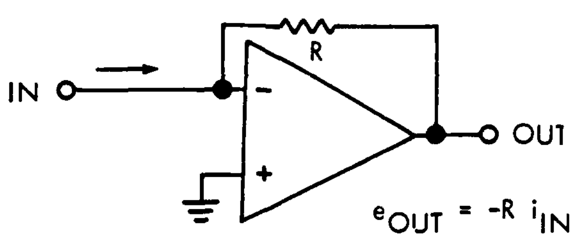
16a. IMPEDANCE MAGNIFIER

$$Z_{IN} = Z_6 \left(\frac{1}{1 + \frac{R_4}{R_3}} \right) = a \; Z_6 \; ; \; a = \frac{R_3}{R_3 + R_4}$$

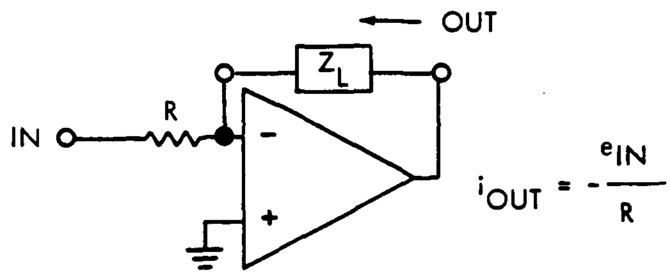


16b. IMPEDANCE REDUCER

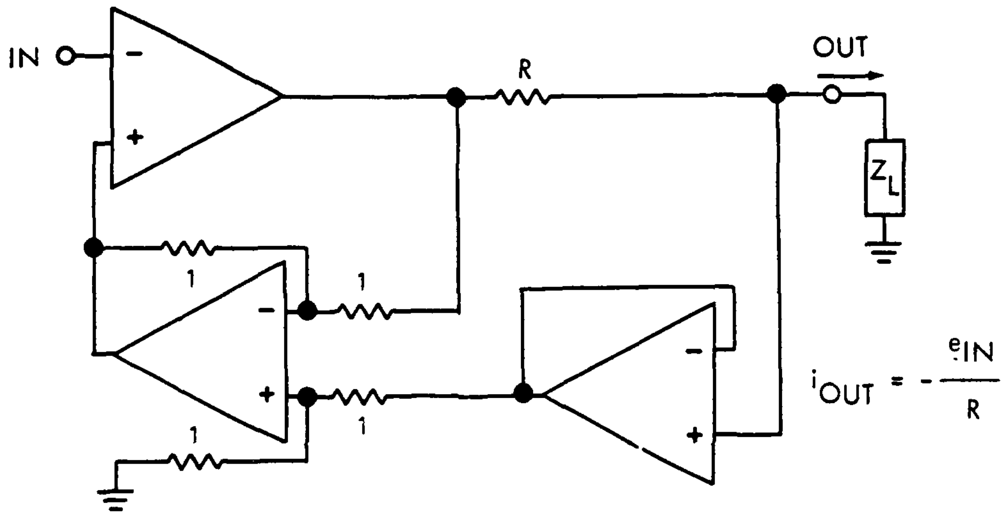
FIGURE 16. IMPEDANCE MULTIPLIERS



17a. TRANSRESISTANCE AMPLIFIER



17b. TRANSCONDUCTANCE AMPLIFIER



17c. TRANSCONDUCTANCE, LOAD GROUNDED

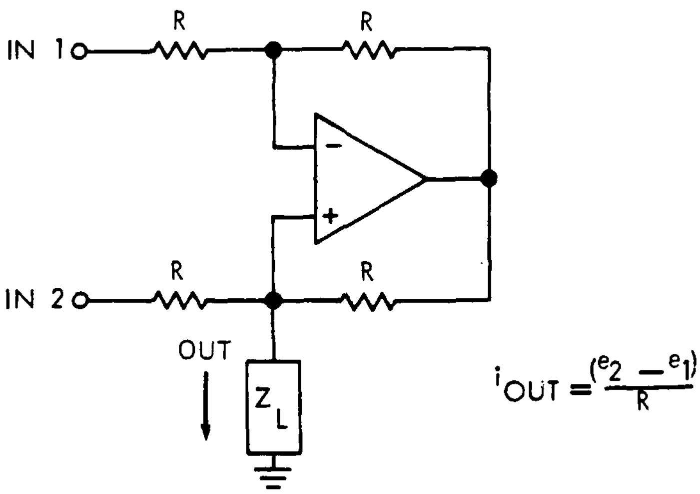
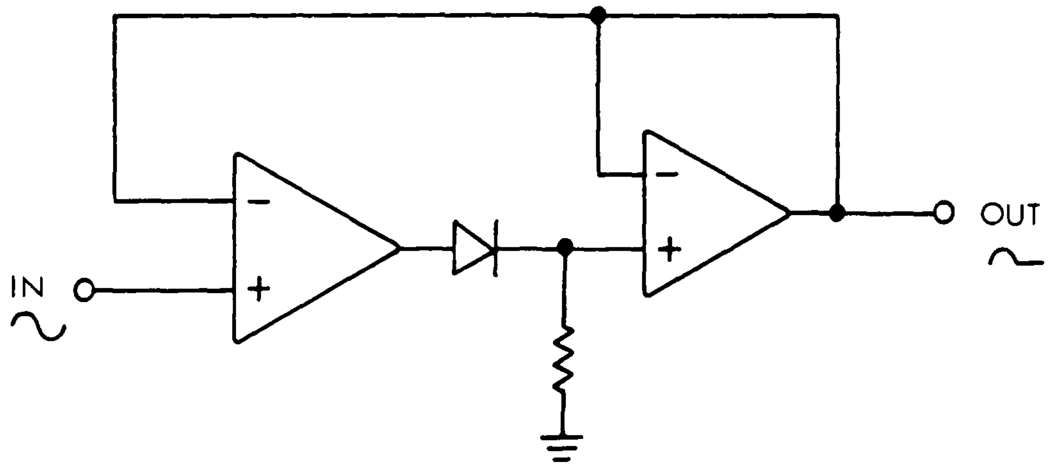
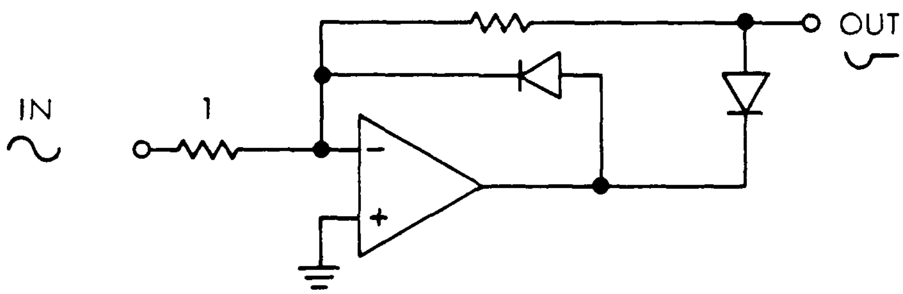


FIGURE 17. TRANS-IMPEDANCE CIRCUITS

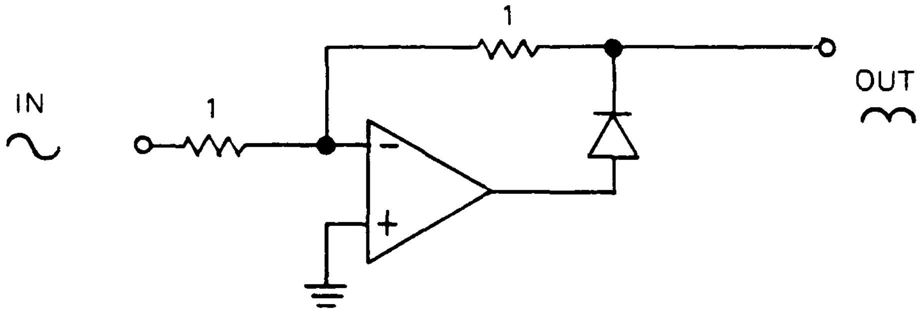


18a. NONINVERTING HALF-WAVE RECTIFIER

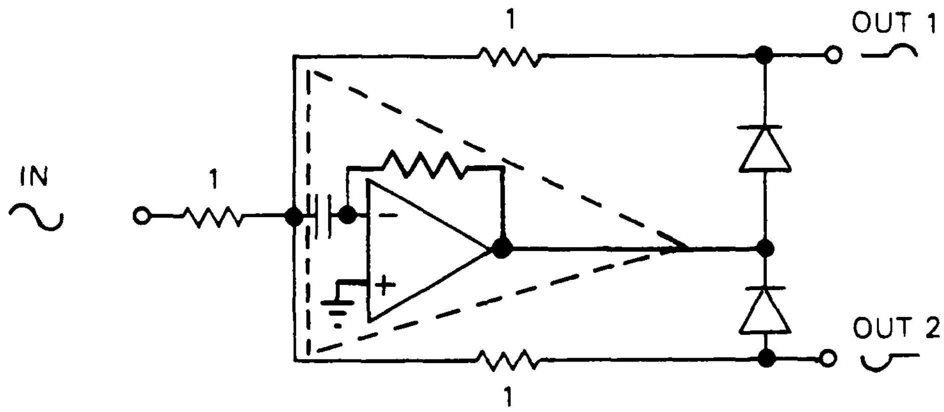


18b. INVERTING HALF-WAVE RECTIFIER

FIGURE 18. RECTIFIERS

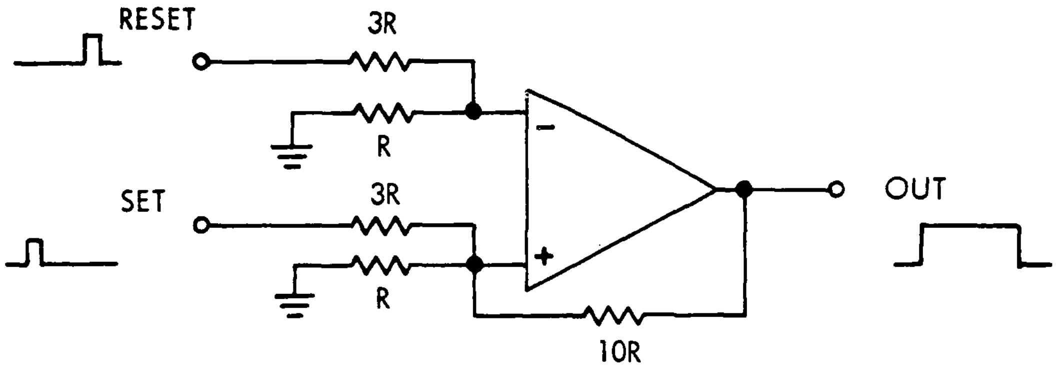


18c. "CHEAP-AND-DIRTY" FULL-WAVE RECTIFIER

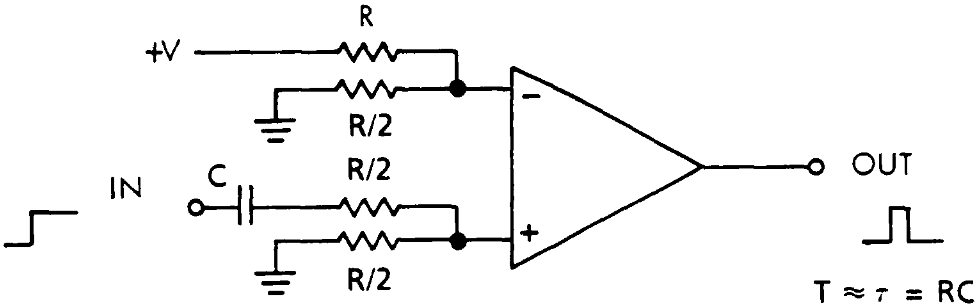


18d. AC RECTIFIER

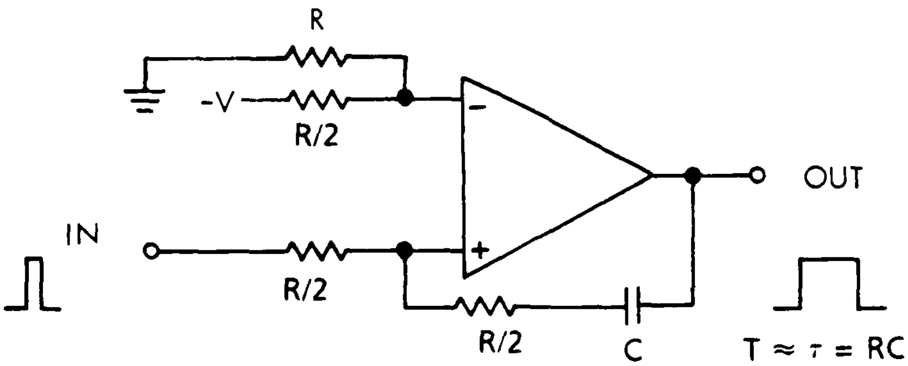
FIGURE 18. (Cont.)



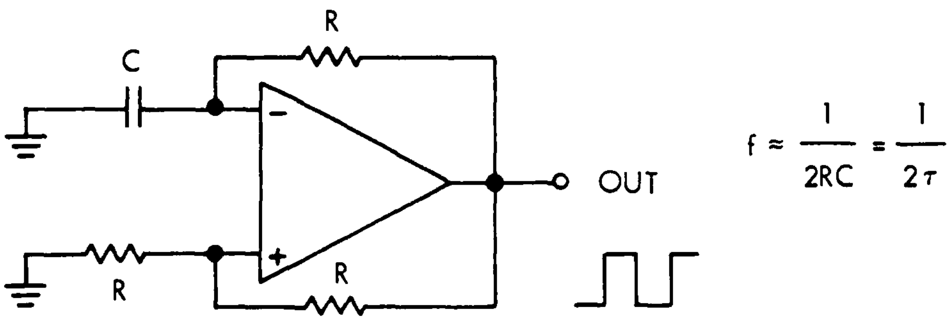
19a. DC SET-RESET FLIP-FLOP



19b. PULSE GENERATOR

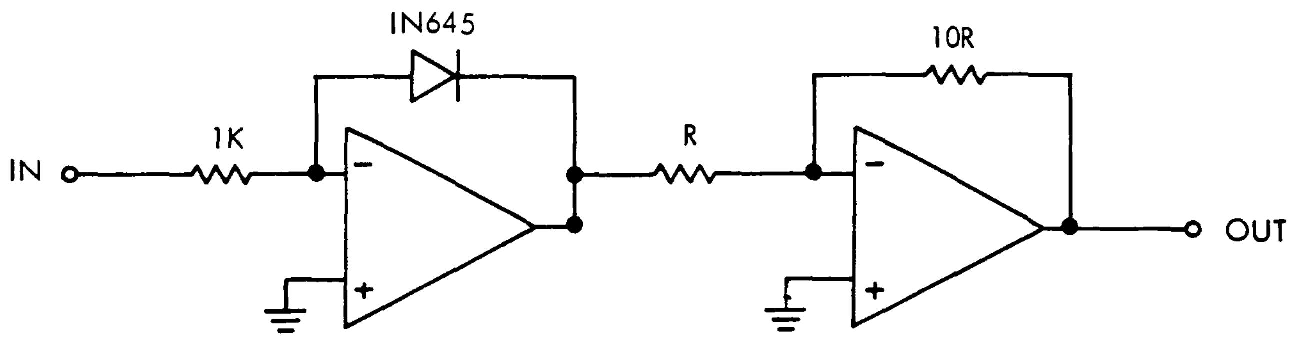


19c. DELAY MULTIVIBRATOR



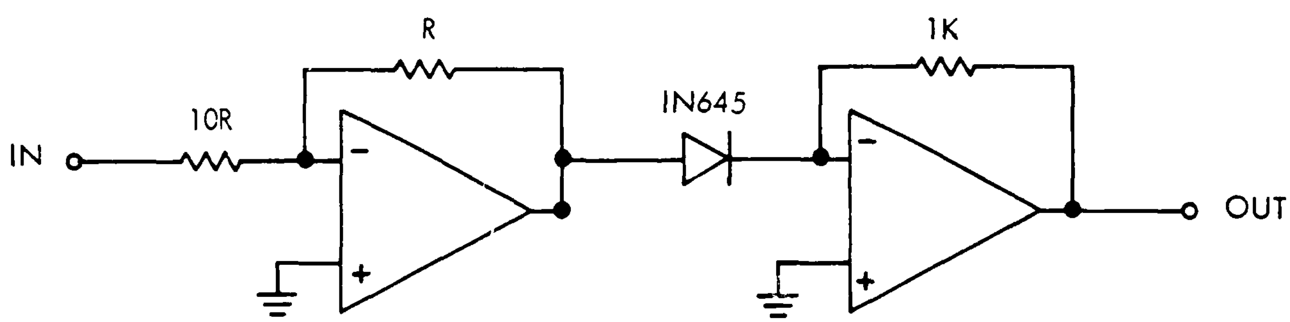
19d. ASTABLE MULTIVIBRATOR

FIGURE 19. DIGITAL CIRCUITS



V_{OUT} (VOLTS) = LOG_{10} V_{IN} (MICROVOLTS)

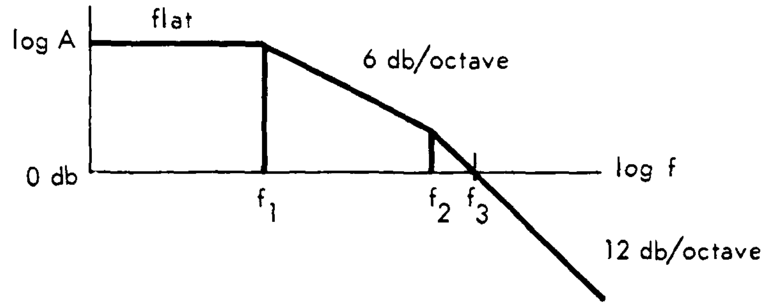
20a. LOG CONVERTER



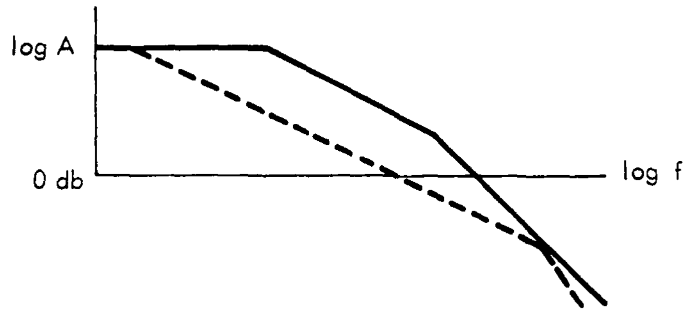
V_{OUT} (MICROVOLTS) = $ANTILOG_{10}$ V_{IN} (VOLTS)

20b. ANTILOG CONVERTER

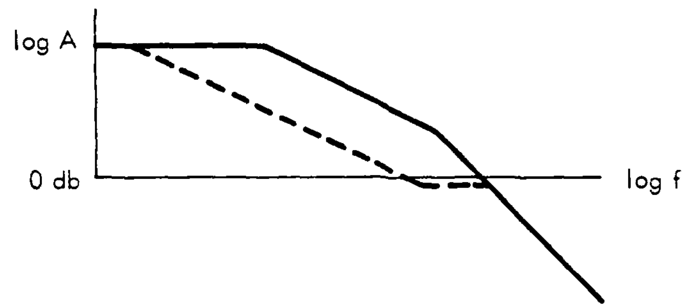
FIGURE 20. LOGARITHMIC CIRCUITS



21a. NO COMPENSATION



21b. LAG COMPENSATION



21c. LEAD-LAG COMPENSATION

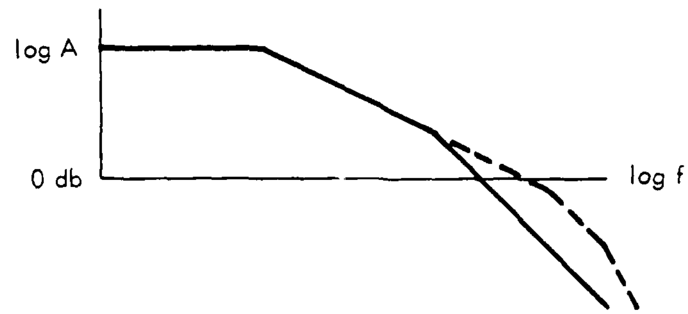
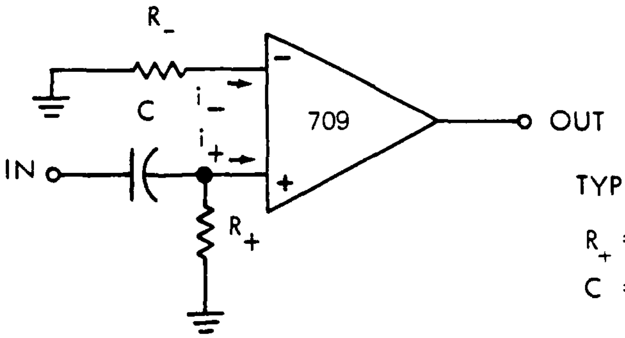


FIGURE 21. OPEN-LOOP CHARACTERISTICS

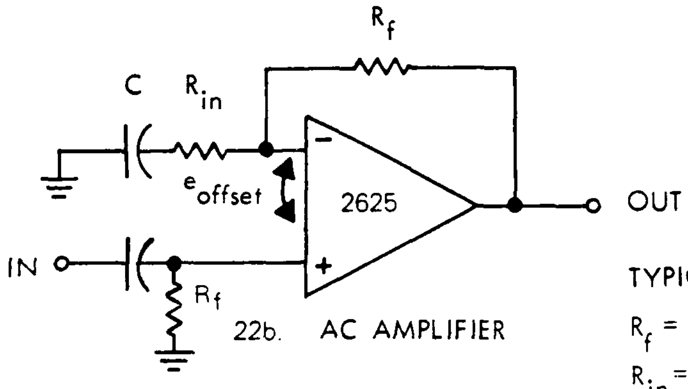


TYPICAL VALUES

$R_+ = R_- = 10K$

$C = 1 \mu f$

22a. AC CLIPPER



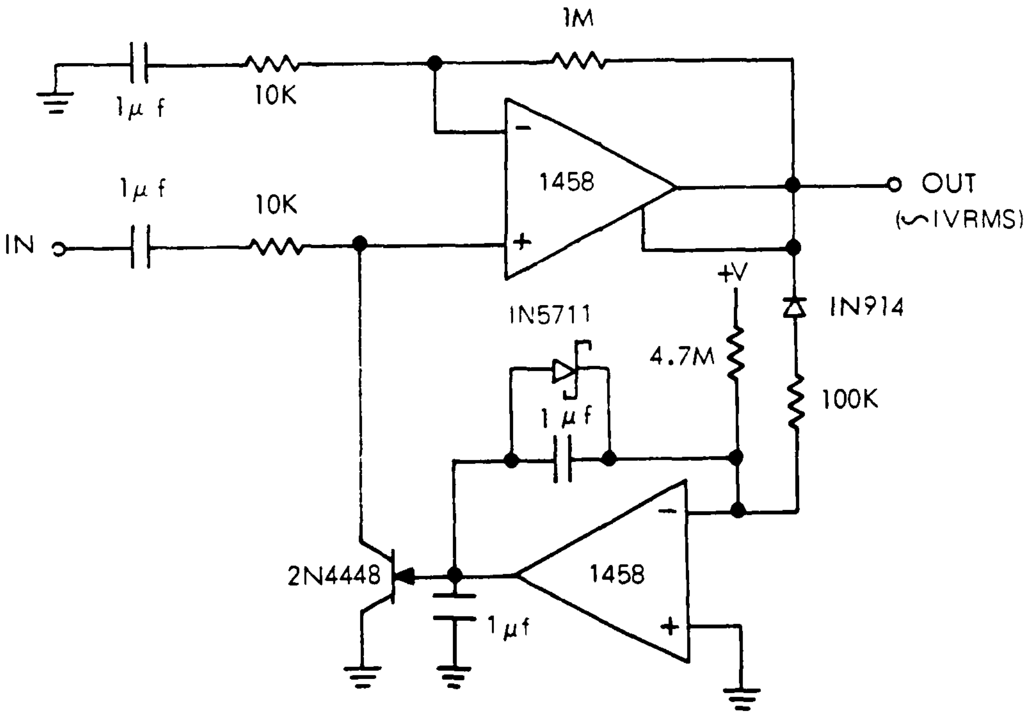
22b. AC AMPLIFIER

TYPICAL VALUES

$R_f = 1\ M$

$R_{in} = 1\ K$

$C = 1 \mu f$



22c. AGC AMPLIFIER

FIGURE 22. DEMONSTRATION CIRCUITS

TABLE 1. MAJOR OPERATIONAL AMPLIFIER SPECIFICATIONS

Specification	Importance
1. Input Impedance (a) Differential (b) Common-mode	Usually not a problem
2. Input Current Offset	Not a problem except high impedance circuits
3. Input Current Offset	Use to advantage
4. Input Voltage Offset	Problem in high DC gain circuits
5. Input Voltage Range	Problem in followers
6. Differential Input Voltage	Problem in clippers
7. Output Current Drive	Limits possible loading
8. Output Resistance	Usually not a problem
9. Output Voltage Swing	Usually limited only a high frequency
10. Common Mode Rejection Ratio	Usually not a problem
11. Supply Voltage Rejection Ratio	Usually not a problem
12. Voltage Gain	Problem at high gain, high frequency
13. Frequency Response	Problem in high speed
14. Slew Rate	Problem in high speed
15. Bandwidth	Indication of frequency response
16. Settling Time	Problem in fast, high accuracy DC circuits
17. Input Noise	Problem in low-level, high-gain circuits
18. Supply Current	Usually not a problem
19. Temperature Coefficients	Worsens certain parameters
20. Other Maximum Ratings	Usually not a problem
21. Price	Usually not a problem

TABLE 2. IC OP-AMPS IN USE IN CODE U25

TYPE	MFR	DESCRIPTION	COMMENTS
μA 702	Fairchild	Original IC op-amp	Unprotected, obsolete
μA 712		Improved version	
μA 709	Fairchild	Original general-purpose	Unprotected, obsolete
μA 715	Fairchild	Original high-speed	Unprotected, obsolete
μA 725	Fairchild	Original "instrumentation"*	Unprotected, obsolete
μA 735	Fairchild	Original "micropower"	Unprotected, obsolete
μA 739	Fairchild	Low-noise dual	Intended for preamp use
μA 749			Different pinouts
LM 381	National		
μA 740	Fairchild	Original JFET input	Obsolete
μA 741	Fairchild	Original compensated	Most common op-amp
μA 747	Fairchild	Dual 741	Obsolete, undesirable pinout
μA 748	Fairchild	Uncompensated 741	
μA 771	Fairchild	FET in; single, dual, quad	Quad discontinued
μA 772			
μA 774			
μA 776	Fairchild	"Programmable" (biasable)	
LM 4250	National		
MC 3476	Motorola		
μA 791	Fairchild	Power op-amp, 1A	Discontinued
LM 101	National	Comparable to 741	
LM 101		Comparable to 748	
LM 102	National	Voltage-follower only	Obsolete
LM 110		Improved version	
LM 107	National	Original low-input current	Obsolete
LM 108	National	Original very-low-input current	Diodes across inputs
LM 112	National	Micropower Superbeta	
LM 116	National	Original Darlington input	Obsolete
LM 118	National	Original compensated high-speed	Diodes across inputs
LM 124	National	Quad 741	

*misleading terminology

TABLE 2. (Cont.)

TYPE	MFR	DESCRIPTION	COMMENTS
LM 148	National		
LM 2902	National		
MC 3503	Motorola		
MC 4741	Motorola		
LM 149	National		Uncompensated
LM 1900	National	Quad "Norton" current amp	Intended for automotive use
MC 3401	Motorola		
LM 192	National	Op-amp/comparator	
LM 13600	National	Dual transconductance	
LF 155	National	JFET input family	Low-power
LF 156			General-purpose
LF 157			Decompensated
LF 347	National	FET input family	Quad
LF 351			Single
LF 353			Dual
LF 400	National	Fast settling FET	
LF 13741	National	741 with JFET input added	
LM 10	National	Op-amp/reference "new generation"	
LM 11	National	Low-offset	
LM 12	National	Power op-amp, 10A	
MC 1529	Motorola	General Purpose	Discontinued
MC 1530			Obsolete
MC 1539	Motorola	Original high slew rate	Obsolete, diodes across inputs
MC 1556	Motorola	Improved 741	
MC 1558	Motorola	Dual 741	
LM 158	National		
LM 2904	National		
μA 798	Fairchild		
MC 1560	Motorola	General Purpose	Discontinued
MC 1590			
MC 3503	Motorola	Quad "single-supply"	

TABLE 2. (Cont.)

TYPE	MFR	DESCRIPTION	COMMENTS
MC 3505	Motorola	Dual Op-amp/comparator	
MC 3571	Motorola	Quad JFET input	
MC 4202	Motorola	Quad programmable (biasable)	Different pinouts
LM 146	National		One version discontinued
MC 14573	Motorola	Quad Programmable CMOS	
MC 34074	Motorola	Quad Fast 741 single-supply	
MC 34084	Motorola	Quad	
MC 34085			
HA 909	Harris	Original low-noise	Obsolete
HA 911		Improved version	
HA 2000	Harris		Discontinued
HA 2060	Harris		Discontinued
HA 2400	Harris	"Programmable" (selective) quad	
HA 2507	Harris	Economy Hi-slew-rate	Compensated
HA 2527			Decompensated
HA 2520	Harris		Uncompensated
HA 2530	Harris	High slew rate, wideband	Inverting only
HA 2541	Harris	Fast compensated	
HA 2620	Harris	Original decompensated wideband	
HA 2700	Harris	Low-Power	Discontinued
HA 2900	Harris	Chopper-Stabilized	
HA 4602	Harris	Quad wideband	Compensated, discontinued
HA 4622			Decompensated, discontinued
HA 5112	Harris	Low-noise dual	
HA 5190	Harris	Wideband	
CA 3010	RCA	General Purpose family	Obsolete
CA 3015			
CA 3030			
CA 3037			
CA 3060	RCA	"Transconductance" amplifiers	Triple
CA 3080			Single
CA 3100	RCA	Wideband	

TABLE 2. (Cont.)

TYPE	MFR	DESCRIPTION	COMMENTS
CA 3130	RCA	Original MOSFET input	CMOS output, uncompensated
CA 3140			Bipolar output, compensated
CA 3160			CMOS output, compensated
CA 3240			Dual 3140
CA 3193	RCA	Low offset voltage	
CA 3440	RCA	"Nanopower" programmable	Discontinued
RM 4136	Raytheon	Original quad 741	Undesirable pinout
TL 061	Texas Inst.	JFET input family	singles, duals, quads, compensated
TL 062			uncompensated, general-purpose,
TL 064			low-noise, low-power, various pinouts
TL 071			
TL 072			
TL 074			
TL 075			
TL 080			
TL 081			
TL 082			
TL 083			
TL 084			
TL 091	Texas Inst.	NFET input family, single-supply	Discontinued
TL 092			
TL 094			
ICL 761X	Intersil	CMOS Low-power family	Single
ICL 762X			Dual
ICL 763X			Triple
ICL 764X			Quad
ICL 7650	Intersil	Chopper stabilized	

TABLE 2. (Cont.)

TYPE	MFR	DESCRIPTION	COMMENTS
OPA 111	Burr-Brown	Low-noise FET	
OP 27	Precision	Low-noise	Compensated
OP 37			Uncompensated
LT 1028	Linear Tech	Ultra-low-noise	
NE 5539	Signetics	Ultra-fast; gigahertz bandwidth	

TABLE 3. TROUBLESHOOTING PROCEDURES

NONLINEAR CIRCUITS

1. Check for correct IC.
2. Check circuit layout, construction and components.
3. Check supply voltages at the pins.
4. Check voltages at op-amp inputs (signals).
5. Disconnect load.
6. Remove op-amp and recheck circuit.
7. Insert new op-amp.

LINEAR CIRCUITS

1. If operation is nonlinear, see above.
2. Do 1-3 as above.
3. Do 4 above, ground output if possible.
4. Check voltages at other op-amp pins.
5. Do 5-7 above.
6. Noise on Cutput
    a. Check Noise on ground leads
    b. Check Noise on power supply leads
    c. Redo circuit layout.
7. High-Frequency Oscillation
    a. Bypass supply leads
    b. Remove load
    c. Increase compensation
    d. Recheck circuit design
    e. Redo circuit layout.
8. Low-Frequency Oscillation
    a. Recheck circuit design
    b. Redo power supply or leads.

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